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<http://dx.doi.org/doi:10.21954/ou.ro.0000d568>

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An experimental investigation into the
Electromagnetic Compatibility aspects of
High Frequency Power Line Communications

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BEng (Hons)

A theses submitted to the
Open University Faculty of Technology
Discipline of Electronics
For the degree of
Doctor of Philosophy

Appendices have been
excluded from this digital
copy at the request of the
university.

Acknowledgements

The author would like to thank his supervisor Dr John Newbury of the Open University for his friendship, advice and wholehearted support during the whole course of this research program.

Thanks are due to professional colleagues Dr Paul Brown and Martin Welch for their continued support and friendship over many years, and in particular for their practical help during the experimental program.

Also deserving of the author's thanks are Dr Peter Nicholson and Dr Javad Yazdani. As recently qualified PhD students, their support and advice has been enormously useful to the author during the writing of this thesis.

The author would of course like to thank his family for their support and encouragement during the whole of the research program and their patience during its later stages.

Abstract

Power line communications technology, long established for low data rate applications, is now charting new territory with respect to data rates and provided services. This can only be achieved by increasing PLC operating frequencies from the low frequency band (below 148.5 kHz) to the high frequency band (1 MHz and upwards). There is now only one technical barrier to widespread deployment – Electromagnetic Compatibility.

Existing low voltage power networks are optimised for the safe supply of electrical energy. Low voltage cables are often pseudo co-axial in their cross section, but when high frequency signals are coupled onto the network, part of the signal will be radiated. There is therefore a potential for interference to be caused to legitimate users of the radio spectrum.

This thesis, and the experimental program underlying it, seeks to quantify potential problems and to propose mechanisms by which they could be mitigated to the extent that wide scale deployment of PLC networks becomes possible.

The first part of the thesis offers a detailed introduction to the topics of electricity supply networks, power line communications, modulation techniques and electromagnetic compatibility. Existing EMC standards are examined and although some do not directly cover power line communications networks, key principals are drawn for later use in standards development.

The thesis then seeks to examine the mechanisms by which high frequency interference might be caused. Radio propagation modes are discussed and a clear technical distinction is drawn between localised interference from a single PLC network to an individual radio user, and cumulative interference from wide spread deployment of PLC systems. Both such scenarios are examined in detail.

The experimental program is described quantifying radiated signal strength regression from a number of power networks and at a number of operating frequencies within the high frequency band. In this context, signal strength regression is the rate at which electrical field strength reduces with increasing measurement distance.

The experimental setup uses a conventional signal generator to supply single test frequencies of known power spectral density, which are coupled onto a power network. The subsequent radiated signal is received via a conventional antenna and radio receiver at a number of locations surrounding the power network at known distances, and signal regression is derived. The experiment was repeated for a number of different frequencies and at representative urban, suburban and rural locations. Indeed, the experimental technique was evolved over a number of months to allow increased portability of the signal receiving equipment, and hence the number of measurements that could be taken.

From the experimental results, presented both in tabular and graphical format, a number of conclusions can be drawn.

Firstly, based on these results, antenna factors in the order of 85 dB/m can be expected of power line communication networks. It can be concluded that the field strength regression to be anticipated from PLC networks is likely to be significantly below the -20 dB per decade ‘free space’ regression figure that has often been used in interference models. In fact a regression figure of -35 dB/decade is more representative of ground wave propagated interference from PLC networks.

It is also possible to conclude that the adoption of orthogonal frequency division multiplexing as a multi-carrier spectral technique offers specific advantages in EMC terms. Due to its nature, it is possible to apply a frequency ‘mask’ to an OFDM based PLC system. Such a mask might be static, applied on a national or regional basis in order to guarantee non-interference with specific frequencies, for example those used for emergency radio channels. It would also be possible to add a dynamic frequency mask, controllable on each PLC system, to mitigate interference with radio services operating within the PLC operating band.

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Summary of abbreviations

AC	Alternating Current
ADSL	Asymmetric Digital Subscriber Line
AMN	Artificial Mains Network
ASK	Amplitude Shift Keying
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CENELEC	Comité Européen de Normalisation ELECtrotechnique
CEPT	Conférence Européenne des administrations des posts et des Télécommunications
CDMA	Code Division Multiple Access
CISPR	Comité International Spécial des Perturbations Radioélectriques
CPFSK	Continuous Phase Frequency Shift Keying
DC	Direct Current
DSL	Digital Subscriber Line
DSP	Digital Signal Processing
DSSS	Direct Sequence Spread Spectrum
EC	European Community
EMC	ElectroMagnetic Compatability
ETSI	European Telecommunications Standards Institute
EUT	Equipment Under Test
FCC	Federal Communications Commission
FEC	Forward Error Correction
FEK	Frequency Exchange Keying
FHSS	Frequency Hopping Spread Spectrum
FSK	Frequency Shift Keying
GMPSK	Gaussian Minimum Phase Shift Keying
GPS	Global Positioning System
HF	High Frequency
ICOM	Icom Inc, Japan (radio equipment manufacturer)
IF	Intermediate Frequency

ITE	Information Technology Equipment
LV	Low Voltage
LW	Long Wave
MF	Medium Frequency
MFSK	Multiple Frequency Shift Keying
MSK	Minimum Shift Keying
MW	Medium Wave
NAMAS	NAtional Measurement Accreditation Service (UK)
OFDM	Orthogonal Frequency Division Multiplexing
OOK	On Off Keying
PLC	Power Line Communications
PLT	Power Line Telecommuniucation
PSD	Power Spectral Density
PSK	Phase Shift Keying
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency
SNR	Signal to Noise Ratio
VDSL	Very high bit rate DSL
VSWR	Voltage Standing Wave Ratio

Preface

On the 29th of June 1883, a patent application was filed with the United States Patent Office by Thomas Alva Edison of Menlo Park, New Jersey [1]. This patent, entitled simply 'Electric-Lighting System', describes a method whereby electrical lamps connected to a direct current circuit, could be remotely switched on and off, independently of other lamps on the same circuit.

Figure 1 - 'Electric Lighting System' from US Patent 430,934, is reproduced directly from Edison's original document. With reference to Figure 1, a direct current generator (A), supplies electrical power to a two-conductor distribution system (1,2) via a double pole double throw switch (B), allowing the voltage polarity to be reversed. Lamps at F, G and those supplied via H will only be lit when current is supplied with one polarity. In each case, the flow of current is controlled by an early electromagnetic relay, described by Edison as a 'polarized apparatus'. Other lamps connected directly to 1 and 2 will be unaffected by a change in polarity.

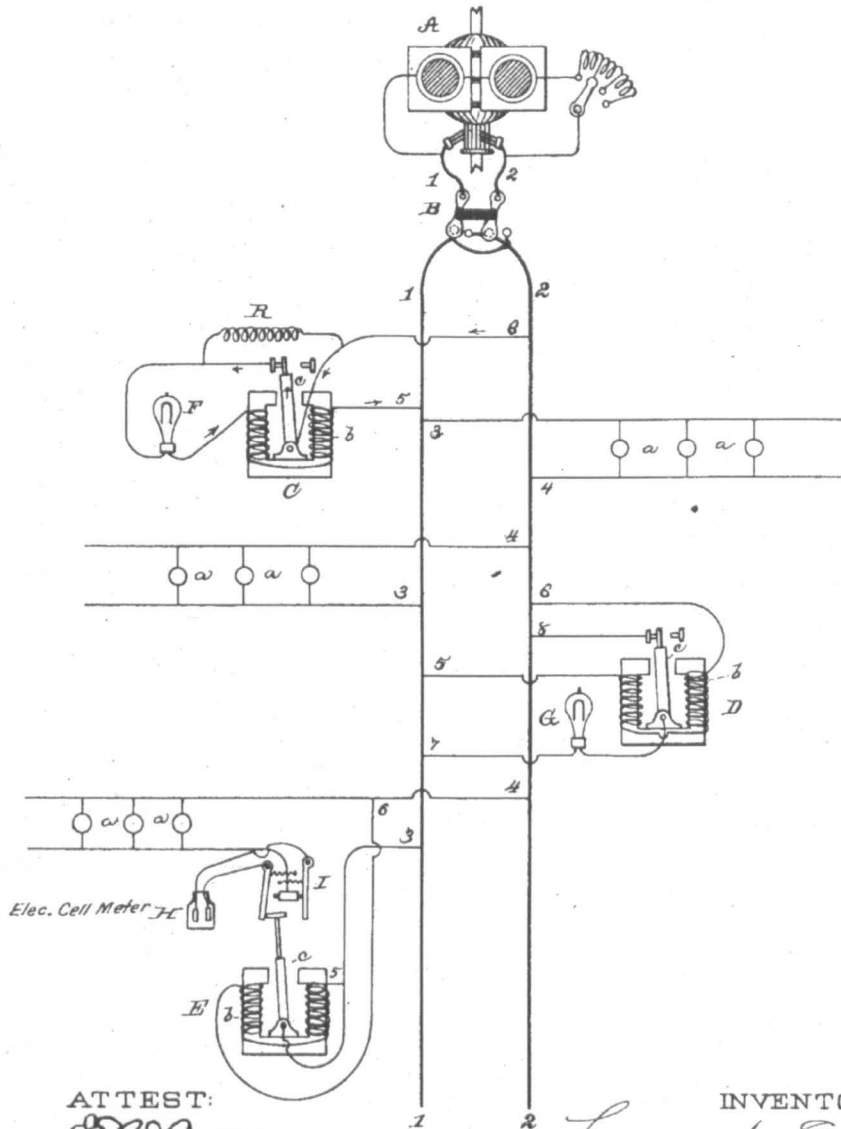
This patent document is perhaps the earliest accessible evidence that a new field of electrical engineering, and thus far a relatively modest one, had begun – Power Line Communications (PLC).

(No Model.)

T. A. EDISON.
ELECTRIC LIGHTING SYSTEM.

No. 430,934.

Patented June 24, 1890.



ATTEST:

O. B. Reynolds
W. L. Lacey

INVENTOR:

Thomas A. Edison.
By Rich. T. Dyer,
Att'y.

THE MORRIS PETERS CO., PHOTO-LITHO., WASHINGTON, D. C.

Figure 1 - 'Electric Lighting System' from US Patent 430,934

1 Chapter 1 – Introduction to Broadband Power Line Communications (PLC)

1.1 The UK electricity supply network

The UK electricity supply current network, although in parts a legacy of technical and investment decisions made 80 years ago, is in fact well suited its task. Figure 2 illustrates its present day configuration [2].

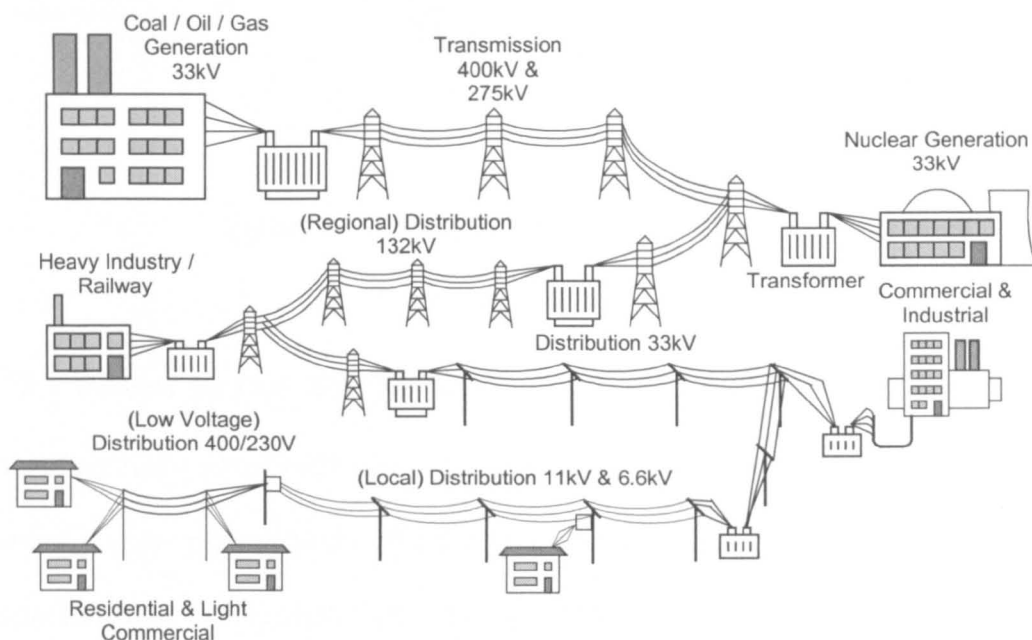


Figure 2 – The UK Electricity Supply Industry

Energy is delivered to consumers at a range of different voltage levels according to their requirements. Available supply voltages range from 132 kV supplies to Heavy Industry and to railways (for transforming to 25 kV single phase traction supplies). Medium and light industry / commercial load can be supplied at 33, 11 or 6.6 kV as appropriate to their needs. Domestic customers are supplied at 230 V, or 400 V if a 3 phase supply is desired. [3]

A typical 230 V domestic wiring arrangement is shown in Figure 3 below.

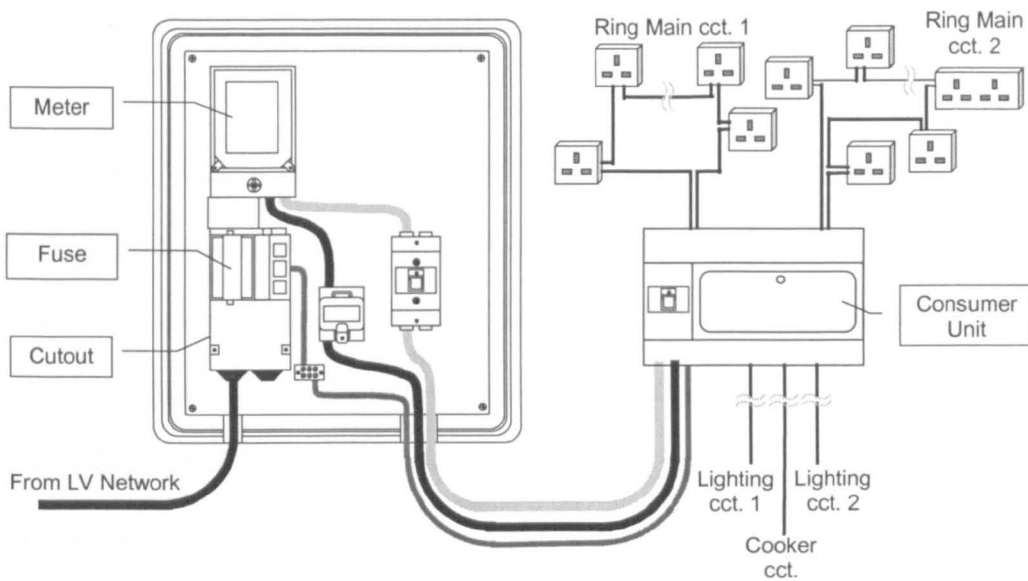


Figure 3 - A Typical UK domestic wiring configuration

1.2 Drivers for PLC Development

The Electricity act of 1989 placed the UK electricity transmission and distribution businesses entirely under private control for the first time in its history. Prior to nationalisation in the late 1940s, it had consisted of a mixture of private supply companies and local authority controlled organisations, interconnected by the developing national grid transmission network. [4]

The newly privatised companies found that profits within their core businesses, the transmission or distribution and sale of electricity, were heavily regulated. As a result, many began to examine possible diversification strategies that could take their technical and commercial expertise into new markets. Telecommunications was soon identified as having considerable potential. [5]

Having operated their own telephony and telemetry networks for many years, in support of the electricity network, a substantial body of telecoms expertise was found to be available.

Additionally, a new entrant to the telecoms market would have no ‘universal service’ obligation and could therefore ‘cherry pick’ the most profitable customers, for example large commercial organisations. It is into this commercial environment that power line communications was introduced in the mid 1990s. Initial systems focused on the provision of telephony services – with the low voltage power network providing a ‘short cut’ local loop and a means of rapidly gaining market share against an incumbent operator. [5] [6]

At the same time the Internet was transforming from a little known academic computer network to the widely adopted network we know today. PLC focus rapidly shifted to the provision of broadband internet access. Here too, incumbent telecoms operators had a competing product – ADSL, but at that time, their service roll out was slow and a perceived ‘window of opportunity’ was available for a fast moving teleco / power company to exploit. [5]

1.3 PLC system types

Modern PLC systems, supporting data rates of 1 MBits and upwards, have evolved to address two distinct markets. ‘In House’ systems make use of a buildings existing electrical wiring to provide a local area network. [7] [8] [9] ‘Access’ PLC systems seek to use the low voltage power distribution network as an alternative to the existing local telephone network connection to individuals properties. This connection is informally known as the ‘last mile’ or ‘local loop’. [10]

1.4 Basic PLC components

In order to realise the potential of a power line communication (PLC) systems, it is necessary to add two basic components to an existing power network – a signal coupling device and a modem.

The coupling unit provides a safe means of coupling telecoms signals onto the power network. Typically this unit is a passive device providing isolation via a capacitive and/or magnetic network. [11] Typical coupling unit designs are discussed later in this chapter.

Having installed two or more coupling units onto a power network, there exists between each pair, a signal path. Although non ideal in terms frequency and phase response, in fact its time and frequency variant characteristics are not unlike a wide bandwidth radio signal path. A cleverly designed modem system can successfully transmit and receive information with a data rate roughly proportional to the available frequency bandwidth.

With in-house PLC systems, the modem and coupling unit are combined into a single physical device, but more often, with PLC access systems, they are installed as separate units, for reasons that are explained fully in Chapter 2 of this thesis.

1.5 Coupling unit design

The device that couples a telecommunications signal onto a normally energised mains network, typically referred to as a coupling unit, fulfils two basic functions. Firstly and most importantly, it provides a safe degree of galvanic isolation between the (non-earthed) mains

conductors and the (non-earthed) telecommunications network conductors. Secondly, the coupling unit provides an efficient bi-directional telecoms signal path, across a required frequency bandwidth, between the telecoms network termination and the mains network. Where necessary, this may include approximate impedance matching. [12]

In fulfilling the above design criteria, it is also important that an installed coupling unit causes no degradation in the reliability of the existing electricity supply network.

Except for a small number of power line carrier systems operating at very low frequency, typically less than 1 kHz, the signal coupling arrangement cannot be purely magnetic. The requirements for non-saturation at low frequency (50 / 60 Hz) and effective response at high frequency (above 1 MHz) are mutually exclusive in magnetic materials. A combined power supply and signal coupling transformer, providing galvanic isolation for both circuits, although a technically elegant solution, is unfortunately not possible for the reasons given above.

Although it is technically possible, purely capacitive signal coupling has a number of serious disadvantages. The X and Y range of capacitors have been developed for use in mains line filters [13]. Both types of capacitors are designed to fail to open circuit, i.e. to provide isolation even after a failure but only the more expensive Y type capacitors are guaranteed to behave in this way. Therefore it is necessary to use a Y type capacitor wherever a capacitor failure would lead to an earthed conductor / enclosure becoming live and risk of electrical shock. In practical terms, all capacitors between Live (Phase) and Earth or Neutral and Earth must be Y type. Capacitors between Live and Neutral or between two

phases of a multi-phase system can be X type, since a failure would not increase the risk of electrical shock.

The sub-classification of these capacitors by number X1, X2, Y1, and Y2, refers to the intended application of the capacitor. X1 capacitors are impulse tested up to 4000 volts and to be installed in a 440 / 250 V low voltage distribution environments whereas X2 capacitors are impulse tested up to 2500 volts and are designed for domestic (250 V only) environments. Y1 and Y2 capacitors are designed for the same environments above but are impulse tested to 8000 V and 5000 V respectively.

For coupling unit design, a simple capacitive circuit arrangement offers no 'fuse blowing' path, with which to protect the power network from closed circuit capacitor failure. Depending on the degree of isolation required for the telecoms circuit conductors, and the requirement to provide isolation under single component failure conditions, it may also be necessary to use two independent capacitors in series to couple onto each phase conductor.

A simple and practical circuit design for a low voltage (LV) residential/commercial/light industrial coupling unit, for operating frequencies of 1 MHz and above, is shown in Figure 1 below.

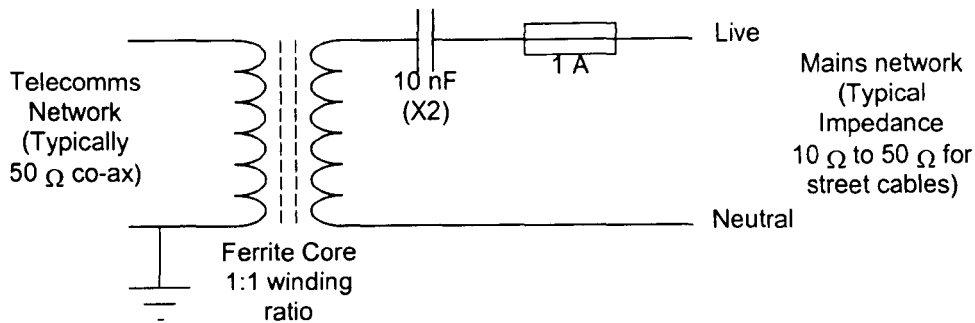


Figure 4 – A simple LV coupling circuit arrangement

This circuit employs an X2 capacitor to block 50 / 60 Hz current, preventing transformer core saturation. A 1 amp fuse, with sufficient voltage rating, protects the power network from close circuit component failure. The transformer is wound onto a ferrite ring core with, in this case, a 1:1 winding ratio, providing no impedance translation between the telecoms and power networks. Other winding ratios can be used to more closely match the estimated impedance of the mains network at the signal coupling point. Galvanic isolation is provided entirely by the transformer.

A modified LV design, for applications where 2 or more phase conductors are present, is shown in Figure 5 below.

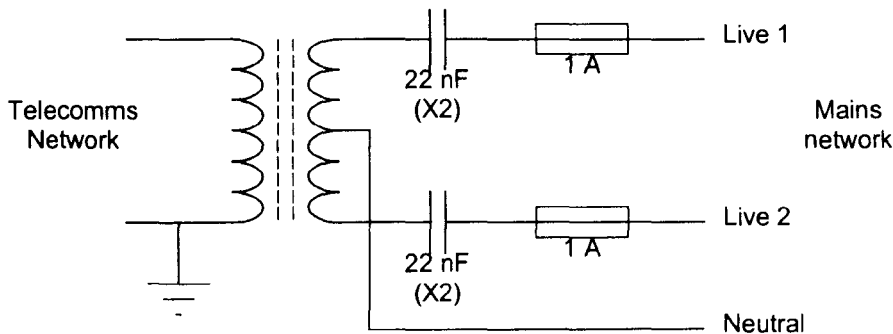


Figure 5 – A dual phase / neutral LV coupling circuit arrangement

A centre tap in the mains side transformer winding provides the necessary fuse-blowing path from each phase to neutral. In this arrangement the transformer is acting as a balun, translating between the power line signal path, balanced about the neutral conductor, and the unbalanced co-ax signal path within the telecoms network.

By coupling onto the mains network in a balanced rather than unbalanced mode, two advantages can be realised. Firstly, the immunity of the signal path to radiated noise ingress will be improved. Secondly, and within the context of power line telecommunications – more importantly, radiated emissions from the balanced power line network will be below the equivalent levels for unbalanced power line network.

The level of balance achievable, and therefore the degree to which these advantages can be realised will depend on the exact cable types and network construction details. The author's previous experience would suggest that where both of the chosen phase pair conductors lie within the same physical cable and where cable branches are minimal, then achievable balance is very good. Where phase conductors run separately, and/or there are several cable branches, the effectiveness of balanced coupling is significantly reduced.

The choice of balanced / unbalanced coupling is often dictated by the existing mains network configuration. Where low voltage live/earth (unbalanced) coupling is desired, the following design Figure 6 may be used.

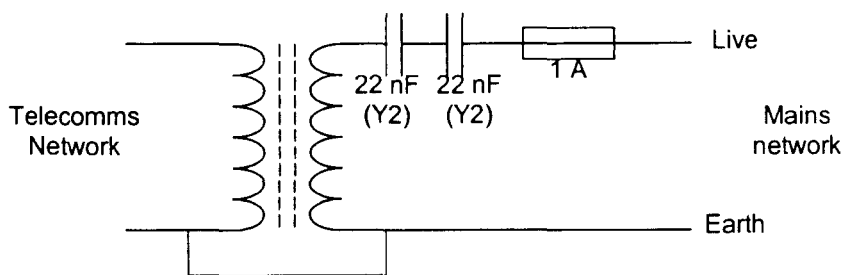


Figure 6 – Live / Earth LV coupling circuit arrangement

In this design the transformer can no longer provide isolation in the event of a failure of the earth conductor. It is therefore necessary to use two Y2 class capacitors in order to insure galvanic isolation under single component failure conditions.

In some cases, the choice of coupling configuration, (Live/Neutral, Live/Live, Live/Earth), will be dictated by the existing power network configuration. For example modern installation practice in the UK requires Protective Multiple Earthing (PME). With a PME installation, the consumer is supplied with live and combined neutral/earth conductors. US installations commonly provide two phase conductors, in a split phase configuration, along with a combined neutral/earth. For this configuration, a suitable coupling circuit is shown in Figure 7 below.

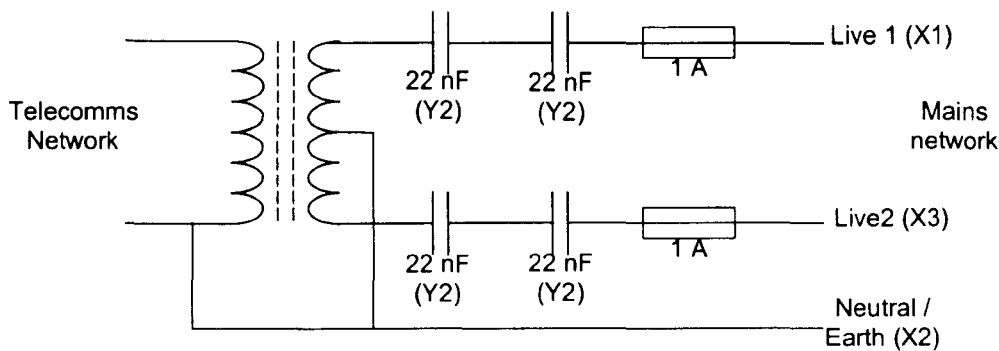


Figure 7 – Dual Phase / Earth LV coupling circuit arrangement

A typical installation arrangement for the coupling unit shown in Figure 8, this time onto a North American style 240 / 120 V network intended for use as a PLT access system. It is necessary, to ensure minimum signal attenuation between the LV supply network and the PLT modem, to connect the coupling unit onto the network side of the meter. Since the first up-stream fuse from this position is often on the MV winding of pad or pole mounted transformer, the need for separate fuse protection within the coupling unit is clear.

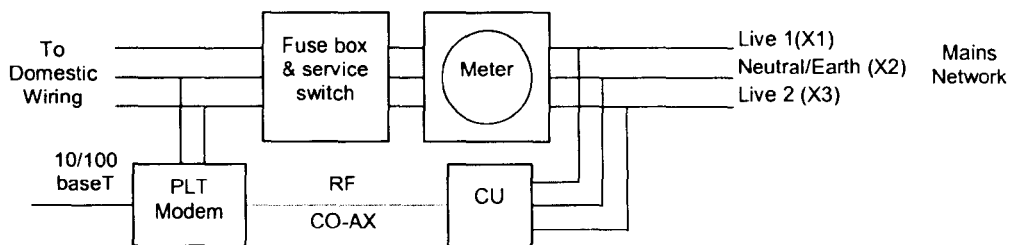


Figure 8 – Installation arrangement at customer premises

In this case, the PLT modem, which draws electrical power from the consumer wiring side of the fuse box and meter, is combined with a 10/100 baseT Ethernet interface, allowing networking via a standard Ethernet network. At the transformer, the LV coupling unit similarly couples the PLT signal onto both phases in balanced mode about the neutral / earth.

1.6 PLC system types

1.6.1 *Simplex (broadcast) systems.*

With a simplex type PLC technology, information is ‘broadcast’ from a single transmitting point on the power network, to all receiving points, by direct manipulation of the power signal. This manipulation generally takes place at the source of supply.

Edison’s system [1], discussed in the preface, is an early example of a simplex type PLC technology. It exhibits interesting technical parallels with more conventional simplex PLC systems. In this case, the polarity of a direct current (DC) supply, manipulated at the source of generation, is used to convey a simple on/off message to every point on the electrical network. Thus, the message can be ‘de-coded’ at each receiving station by the operation, or non-operation of a relay.

Simplex PLC systems are often used where selective load switching, for example of lighting or heating devices, is required. The X10 home automation system [14] has enjoyed considerable success, within the consumer market, with a simplex PLC design. The X10 home automation system was developed in the United States during the 1970s. It uses a simplex transmission system based on zero crossing point manipulation (covered later in this chapter) to control up to 256 appliances or lights from a single remote control. More recent developments have allowed computer control of the appliances allowing devices to be switched on and off according to a pre-arranged schedule or in response to specific computer detected events.

Clearly, the receiver in a simplex PLC system has no means of message receipt, or of requesting re-transmission. In simple, non-critical applications, this problem can be overcome by repeating the message until, to an acceptable level of probability, it will have been received without errors.

1.6.2 Duplex & Half Duplex systems.

As the complexity of transmitted messages, and therefore the scope for error increases, it becomes necessary to implement half-duplex or duplex PLC systems. With this type of PLC system, message receipt can be acknowledged and re-transmission requested, allowing a more robust system to be designed.

1.7 PLC signalling methods and modulation techniques.

Although DC polarity switching is useful as an example of a simple PLC system, the universal adoption of alternating current (AC) power distribution systems limits its practical applications. For AC systems, a number of other techniques have been developed.

1.7.1 Zero Crossing Point Manipulation.

This method of PLC signalling transmits binary messages as a series of electrical pulses coupled onto the mains network, often at the source of supply. Practical measurements have shown that the time during which the AC waveform is close to zero volts, i.e. its zero crossing point, is the optimum time for pulse transmission and reception. During this time, electrical noise caused by other devices on the network is at a minimum, thus the probability

of successful pulse detection is increased and the probability of incorrect detection is decreased. Additionally, the effect of a noise spike at this point in the waveform, on overall power quality, is minimised.

Power quality is essentially a measure of the undesired harmonic content, and frequency of noise spikes on an AC power distribution system. A number of standards now exist, at national and international levels, by which it can be defined. The increasing use of sensitive electronic devices, powered from AC power networks, makes power quality an increasingly important issue for potential PLC systems.

Synchronisation between the transmitting and receiving units can be achieved using the 50 / 60 Hz power signal present at each location. Figure 9 shows a single byte of binary data, transmitted using this method of PLC signalling.

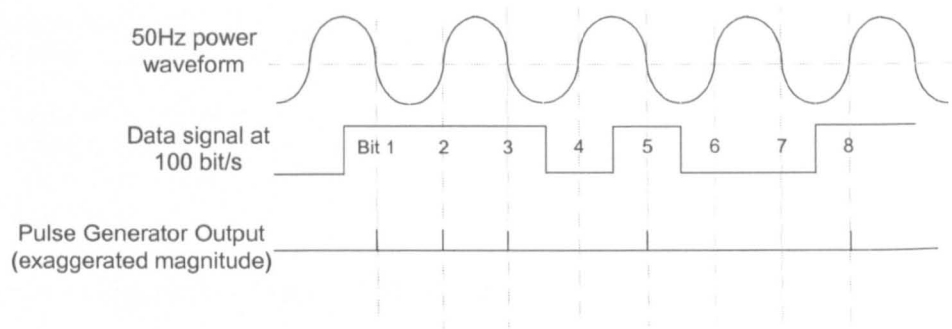


Figure 9 - Zero Crossing point manipulation

Despite its spectral in-efficiency and inherently limited data rate, zero crossing point manipulation has been implemented in a number of PLC systems.

For higher data rate PLC applications, it becomes necessary to transmit information using some form of modulated carrier signal. A number of modulation methods are available, each having distinctive technical advantages and disadvantages along with varying levels of implementation complexity.

1.7.2 Amplitude Shift Keying (ASK)

This technique is similar in many respects to amplitude modulation. Binary digits, in modulation terms often referred to as mark (1), and space (0), are represented by a single carrier signal transmitted at one of two different amplitudes. The most common form of ASK, where one of these amplitude levels is zero, is often referred to as On Off Keying (OOK). Figure 10, below, shows a data signal along with the corresponding OOK modulated signal in both time and frequency domains.

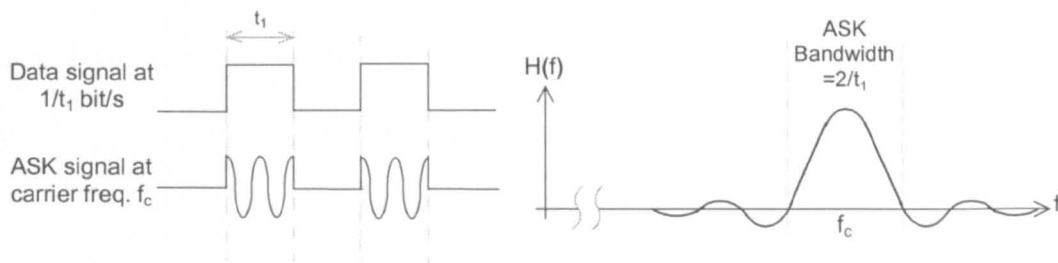


Figure 10 - Typical ASK (OOK) signal and spectral envelope

In Figure 10, above [40] [15], the data rate is $1/t_1$ bit/s, The symbol or baud rate, i.e. the maximum possible number of signal state transitions (from on to off and vice versa), is equal to the data rate. The modulation scheme is said to have an efficiency of 1 bit/baud. With other modulation schemes, most notably some implementations of phase shift keying (PSK), the baud rate and data rate are not necessarily equal.

The minimum bandwidth required to implement an ASK type system is equal to twice the data bandwidth. For example, a system with a modest 128 kbit/s overall data rate, i.e. 131,072 bit/s, would occupy at least 262 kHz of spectral bandwidth.

When the data signal is applied to the modulator in an unmodified state, i.e. with sharp transitions between successive digits, the ASK spectra will extend well beyond twice the data bandwidth, as shown in Figure 10. Band pass filtering within the transmitter at the Intermediate Frequency (IF) or transmitted frequency stage can reduce this problem. A more flexible solution is provided by data pulse shaping - passing the data signal through a low pass filter, before modulation.

In order for both the sending and receiving device to retain synchronisation, it is necessary to avoid long sequences of '0's or '1's. In some cases, this can be achieved through the transmission of short data packets, or individual bytes, each containing a mandatory preamble section for synchronisation. In other cases, more complicated procedures such as 'return to zero' or bi-phase (Manchester) coding could be employed, at the expense of increased ASK bandwidth.

OOK modulation has been successfully used on low frequency PLC systems, where its relative simplicity makes it an attractive option.

1.7.3 Frequency Shift Keying (FSK)

In many respects, this technique is the binary equivalent of frequency modulation. Successive binary digits are represented by the presence of one or other of two carrier

frequencies. By convention the higher frequency represents a binary '1', or mark, and the lower frequency represents a '0' or space [15]. Figure 11, below [40] shows the generated carrier signals, along with the FSK modulated signal in both the time and frequency domains.

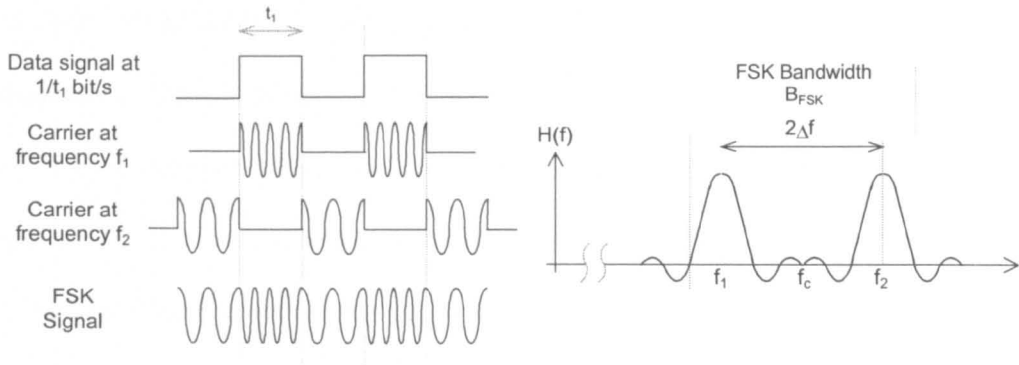


Figure 11 - Typical FSK signal and spectral envelope

By considering the FSK signal to be the output of a standard analogue FM modulator, it is possible to derive a carrier frequency, $f_c = f_1 + (f_2 - f_1)/2$, and a carrier deviation, $\Delta f = (f_2 - f_1)/2$. Thus the modulation index becomes $\beta = \Delta f / B$, where B is the data rate in bit/s. The bandwidth occupied by an FSK signal of data rate B , and modulation index β is given by $B_{FSK} = 2B(1 + \beta)$.

For the 128Kbit/s example, with a frequency deviation, Δf , of 131.072 kHz, (matching the data rate B), the total occupied bandwidth would be 524.288 kHz.

In the Figure 11 illustration, the carrier deviation is greater than the data rate. In this instance, as with any FSK modulation where Δf is greater than or equal to B , the spectral envelope is equivalent to two separate ASK signals at f_1 and f_2 . This is generally termed wideband FSK modulation. There is, in theory, no performance advantage for increasing Δf beyond B . In

situations where the communications channel exhibits frequency selective and time variant fading, most notably with long distance HF wireless links, the data signal can quite often be recovered from just one of the two carriers if the frequency deviation is large.

For high data rate applications, Δf can be set at less than B . This is referred to as narrowband FSK modulation, (narrowband referring to the carrier deviation rather than the data rate). In this situation, the observed spectral envelope is somewhat different - the signal power is contained within the carrier, and a number of side-bands spaced at intervals of the data rate ($1/t_1$). As the data rate is increased, the carrier, (conveying no data itself), contains an increasingly larger amount of the total transmitted signal energy [15]. Wideband and narrow band FSK are shown in Figure 12, below [40].

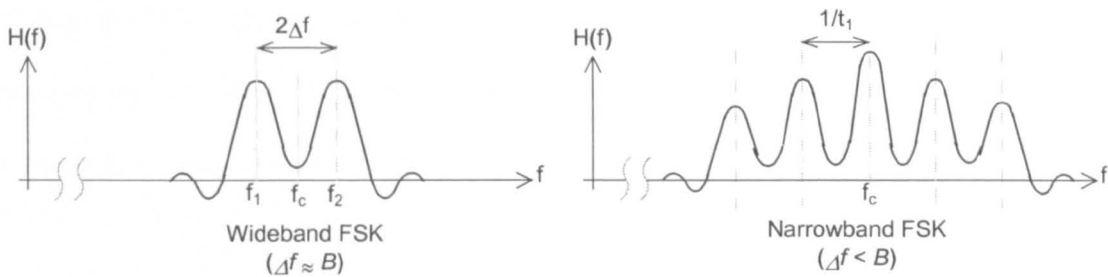


Figure 12 - Wideband and Narrowband FSK compared

FSK can either operate in coherent or non-coherent manner. Within a coherent modulator circuit, and assuming two separate frequency generators are employed, there must be a known phase relationship between each carrier frequency, f_1 and f_2 , the data rate, and a fixed reference.

With a non-coherent FSK modulating circuit, the phase relationship between f_1 and f_2 at the frequency transition is not considered, and is often completely random. This allows a simpler modulation circuit, at the expense of increased 'out of band' interference.

As with ASK, the transmitted signal will occupy bandwidth beyond the theoretical minimum, and again, a number of techniques, in addition to band pass filtering, are available to mitigate this effect.

Data pulse shaping, in FSK terms referred to as Frequency Exchange Keying (FEK) or, in some implementations, Gaussian Frequency Shift Keying (GFSK), is often employed. In this case, the original data signal, used to control f_2 , and its inverse, used to control f_1 , are both manipulated prior to their use by the modulation circuit.

Continuous Phase FSK (CPFSK) can achieve similar results to data pulse shaping by generating the FSK signal in a different way. A frequency lock loop arrangement is employed to generate the (coherent) FSK signal. The feedback loop is arranged in such a way as to slow down frequency transitions until they occupy up to 10% of each symbol time.

Multiple Frequency Shift Keying (MFSK) is an interesting variation to the original FSK technique in which more multiple frequency combinations are employed to represent binary digits. A number of advantages can be realised, including reduced inter-symbol interference, and increased data rate. MFSK should not be confused with another FSK variant Minimum Shift Keying (MSK).

With MSK, a CPFSK variant, a modulation index of 0.5 is used. The deviation, Δf , is therefore half the data rate, B . This is the minimum deviation for which carriers at f_1 and f_2 are orthogonal, hence the term Minimum Shift Keying. Gaussian Minimum Shift Keying, GMSK, is a widely used derivative of this technique offering enhanced spectral efficiency.

With FSK and related variants, the same synchronisation problems as ASK can appear when long strings of '0's or '1's are present within the data stream. These problems can be solved by adopting the same suggested techniques.

1.7.4 Phase Shift Keying (PSK)

With PSK, binary digits are represented by phase shifting the carrier signal. At its most simple level, Binary Phase Shift Keying, two phase shifts are used, usually 0° representing binary 0, and 180° representing binary 1. Figure 13, below shows a typical PSK signal, in this case Binary PSK, along with the observed spectral envelope [40].

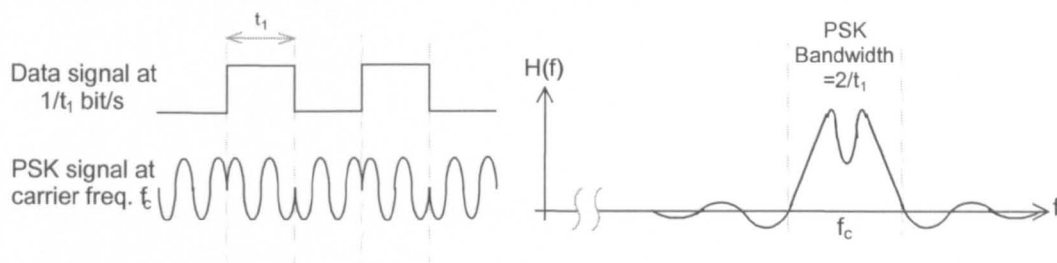


Figure 13 - Binary PSK signal and spectral envelope

Conventional PSK requires the phase of the original carrier, prior to modulation to be known to the receiver. In some applications, the phase information can be recovered from the received signal, although with most communication channels, subtle and time variant phase shifting make this impossible.

Differential PSK overcomes this problem by using the phase of the previous symbol as a reference for the current symbol. Figure 14 compares PSK and Differential PSK for the same data stream [40].

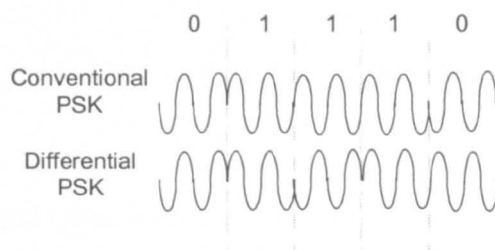


Figure 14 - Conventional and Differential PSK compared

In this example, for Differential PSK, binary 0 is represented by 0° phase shift, from the previous symbol, whereas binary 1 is represented by a 180° phase shift from the previous symbol.

Binary PSK, in common with all other modulation schemes discussed so far, offers only 1 bit / baud. Each symbol - exhibiting either a change or zero change in the carrier's state (amplitude, frequency or phase) represents only a single bit within the data stream. If the number of phase shifts available to the modulator was increased from two to four, it would be possible for each symbol to represent two binary digits, (1 dibit). Thus a modulation efficiency of 2 bits / baud can be achieved without any increase in required bandwidth.

The modulated signal is generally represented by the sum of two quadrature frequencies, i.e. the original carrier, and a $+\pi/2$ phase shifted carrier at the same frequency [40].

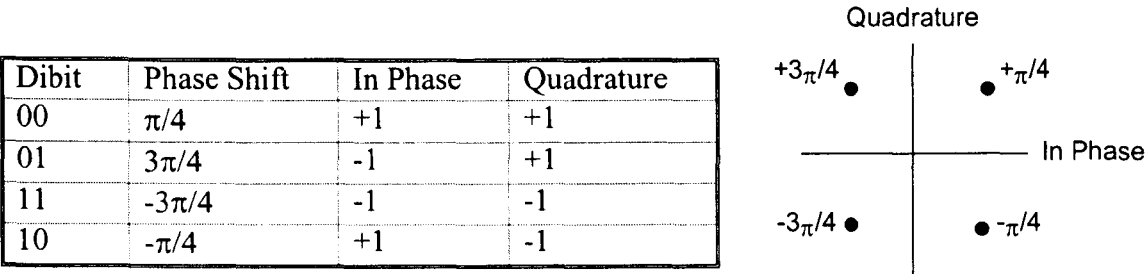


Figure 15 - QPSK phase shifts, corresponding dibits and signal space diagram

Where necessary, Differential QPSK can be utilised just as successfully as standard QPSK.

It is possible to implement PSK systems with 8 or more phase shifts, distributed equally between $\pm\pi$. However, for high bandwidth data communications, a more reliable technique has been developed that combines phase shifts with amplitude variations. Quadrature Amplitude Modulation, QAM, can offer 4 bit / baud performance, by utilising 16 discrete phase/amplitude shifts, shown in Figure 16 [40].

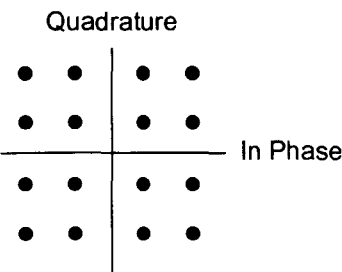


Figure 16 - Signal space diagram for 16QAM

Where communication channel conditions are particularly favourable, 64 level QAM and higher are possible.

QPSK and QAM modulation offer obvious advantages over ASK and FSK, in terms of spectral efficiency. However, in addition to the increased error susceptibility of these PSK variants in the presence of noise, they rely on having a substantially flat channel frequency and phase response - something which no power network can deliver. Without a complex and adaptive channel equalisation capability, and the selection of a part of spectrum with consistently low noise, it would be impossible to use a single xPSK or xQAM carrier for high bandwidth PLC. The potential broadband PLC application of these modulation techniques lies in their use within multiple (low rate) carrier systems - OFDM being an obvious example.

A number of high bandwidth PLC companies have based their technology on PSK type modulation schemes.

1.8 Relative noise performance of ASK, FSK and PSK

When considering noise performance in data communications, it is usually the signal to noise power ratio, (SNR), required for a given bit error rate, (BER), that is discussed. In most cases the noise is assumed to be Gaussian in nature, i.e. to have a completely flat frequency spectrum. For a low frequency powerline communication system (i.e. up to 10 kHz) this would not be a true noise model – the noise would tend to be concentrated in harmonics of the 50 / 60 Hz mains signal. It would also be time variant in nature. The noise amplitude would peak at the time the 50 / 60Hz mains signal was at a maximum, and would be minimised when the mains signal crossed zero. Thus the noise amplitude would follow a 100 / 120Hz waveform. Above 10 kHz the harmonic effect is severely reduced and at higher frequency, 1 MHz and upwards, it is insignificant.

Band limited noise bursts, and impulse type noise events are not modelled, but their relative levels are likely be well above the PLC signal for the duration of the event, preventing any data communication.

By mathematical analysis, it can be shown that BPSK requires the smallest signal to noise ratio, which means that it can be operated over the worst channel conditions. For the same error rate, an FSK system would require an SNR 3dB higher. Similarly, for the same error rate, ASK would require an SNR of 3dB higher again. The above assumes comparison on a peak signal power basis and coherent detection in all cases. [40]

Comparing on a mean power basis, (for OOK the signal power is zero for 50% of symbols), ASK and FSK require the same SNR for a given BER, with BPSK still offering a 3dB SNR advantage. For a BER of 10^{-9} , BPSK requires a mean signal power to noise power ratio of approximately 19dB. Coherently detected ASK and PSK would require approximately 22dB. The quoted BER generally includes the use of forward error correction, (FEC), to detect and correct errors without data re-transmission.

In practice, (non-coherent) envelope detection is often used to decode ASK and FSK signals, further deteriorating their SNR performance. QFSK and QAM also require increased SNRs in order to support given BER. A common approach, among broadband PLC chipset designers, is to select a modulation scheme for each channel dynamically. Once a modest rate digital communication path between two network nodes has been established using ASK, FSK or BPSK, the two nodes will negotiate the most effective modulation scheme, possible QFSK or QAM, on a channel by channel basis. A balance will be found between raw data

rate and BER. Since the channel conditions are dynamic, the modulation schemes may need to be automatically renegotiated on a regular basis.

1.9 Multiple channel techniques including OFDM

It is often advantageous to divide the overall communications channel into a number of narrow frequency bands, and use each band as an independent data channel. The frequency bands may cover the entire communications channel, or may be selected, (dynamically or by pre-determination), from it. Two main advantages are realised by applying this technique.

The communication channel across a narrow frequency band is substantially flat in frequency terms. A conventional modulation technique can be applied without complicated channel equalisation.

Also, narrow band noise can also be confined to only small proportion of channels, allowing data transmission to continue over the remaining channels.

Orthogonal Frequency Division Multiplexing, (OFDM) allocates communication channels such that each carrier is orthogonal with every other carrier - see Figure 17, below [40].

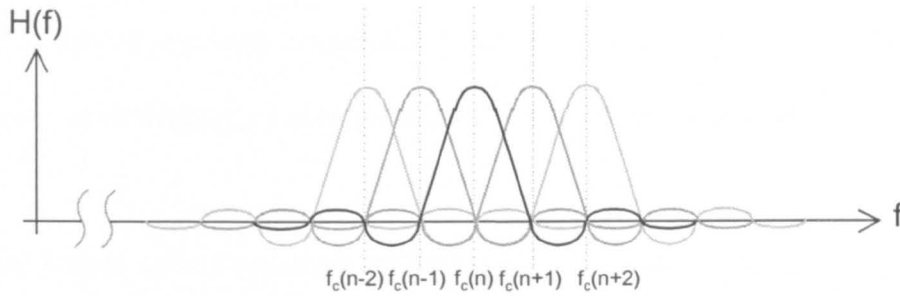


Figure 17 - Frequency spacing of adjacent OFDM channels

For this approach to be successful, the spectral envelope of each modulated carrier must fit within a classic square wave spectral envelope, mathematically described as $H(x) = (\sin x) / x$. ASK and PSK already fit within this spectral envelope, and wideband FSK can be configured such that each carrier does. By arranging each modulated carrier such that the carrier frequency of one carrier falls exactly over a null in the spectral envelope of each other carrier, it is possible to coherently demodulate each carrier without inference from adjacent carriers.

Providing the above conditions can be satisfied, it is entirely possible to apply different modulation schemes to each carrier. This technique has been widely adopted among OFDM based PLC designs where, according to available SNR, individual carrier modulation is dynamically selected from - null (no signal), BPSK, QPSK and QAM [16] [17]. There is generally a degree of frequency flexibility within OFDM systems, since most of the modulation / de-modulation process is implemented in software [18] [19].

1.10 Spread spectrum techniques.

Until recently, spread spectrum communications were used exclusively by the military, in order to prevent eavesdropping and to protect against deliberate signal jamming.

The principal behind spread spectrum technology is to spread the transmitted signal energy over as wide a frequency range as possible. In practice, there are two ways of achieving this.

With Frequency Hopping Spread Spectrum, (FHSS), the carrier frequency of the modulated signal is shifted many times during transmission, according to a pre-determined sequence known to the transmitter and receiver. Where the shifting sequence is sufficiently random within a sufficiently wide frequency range, data communication can take place even if there are a number of 'bad' frequency positions. A 'bad' frequency position may occupy a section of the overall channel exhibiting particularly severe signal attenuation or may suffer from excessive noise, perhaps even deliberate jamming.

The alternative spread spectrum technique is referred to as Direct Sequence Spread Spectrum, (DSSS). In this case, the data signal is XOR'd with a higher rate Pseudo random 'spreading code', prior to modulation using BPSK, QPSK or QAM. This code, known to the transmitter and receiver, is calculated to spread the modulated signal energy, over a wide frequency range.

In most practical cases, the DSSS signal level, measured within a narrowband section of the occupied spectrum, is below the noise floor. Figure 18, shows a typical data stream, spreading code, XOR'd combination and BPSK modulated signal [40].

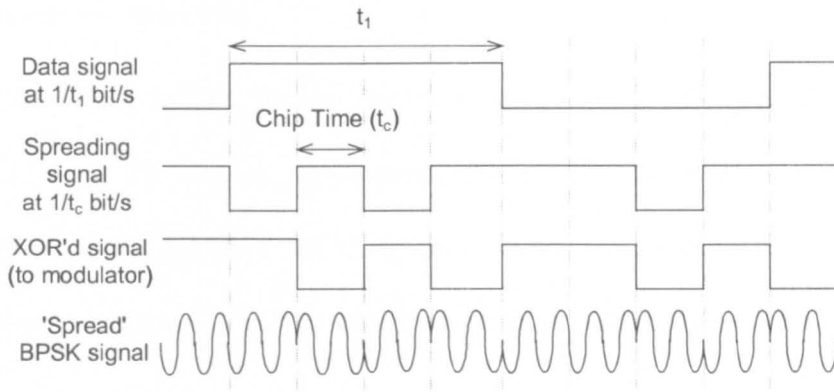


Figure 18 - Direct Sequence Spread Spectrum

It is possible for a wideband receiver to recover to re-generate the original carrier frequency and phase. The original data and spreading code combination can now be de-modulated, and with prior knowledge of the spreading code, the original data can be recovered.

The excellent noise performance of DSSS systems can be used to multiplex many data signals onto the same frequency spectrum. Although there may be two or more transmitters operating at any one time, it is possible to selectively demodulate an individual data signal, or indeed two or more data signals, by using each data signal's unique spreading code. This technique is most commonly referred to as Code Division Multiple Access, (CDMA). [20]

The wide frequency bandwidth required to use DSSS effectively can require increased complexity in both the receiver and transmitter, but the advantages gained in terms of resistance to frequency selective fading and narrowband noise have lead to DSSS recently being adopted for consumer applications. Indeed, the low power spectral density, (PSD), achievable with DSSS, along with its inherent channel adaptation, would appear to offer an attractive scheme for power line communications.

1.11 PLC Frequency Selection

Traditional PLC system designs, for low data rate applications, have operating frequencies in the low kHz. At these frequencies, signal propagation is broadly similar to the power signal itself at 50 / 60Hz. Although the level of electrical noise is high, this can be partly mitigated by avoiding low multiples of the power frequency, where harmonic noise will be at a maximum.

The widespread use of long wave (LW) broadcasting in Europe effectively imposes an upper frequency limit of 148.5 kHz. Similarly, the maximum usable frequency in the United States and Asia is 450kHz – the bottom of the medium wave (MW) broadcasting band. There is a large installed user base of LW and MW receivers, operating in normal domestic and commercial environments, in close proximity to power networks. Thus the selection of a PLC operating frequency within these bands would risk widespread disruption to these radio services.

Within these practical frequency limits, two main PLC standards have emerged. European Cenelec Standard EN50065, itself a more detailed version of CISPR standard IEC61000-2-2, defines frequency selection, transmitted power levels and coupling impedances within the frequency range 3 kHz to 148.5 kHz.

In the United States, where standards are set by the Federal Communications Commission (FCC), one section of FCC part 15 defines permitted PLC transmitted power levels across a much wider frequency range.

A number of PLC designs, successfully fulfilling low and medium data rate applications can demonstrate compliance with one or both of the above standards. However, in order to realise fully the high rate, perhaps even multi-megabit, telecommunications potential of power networks, it is necessary to move away from 'traditional' low frequency systems operating below 450kHz, and consider the use of the high frequency (HF) band.

At HF, between 1 and 30 MHz, power networks offer some of the most challenging communications channels faced by electronic engineers. Problems include frequency selective fading and high levels of white, coloured and impulsive noise. In addition, the network impedance seen at each signal coupling point varies within a relatively wide range and is also frequency selective. Furthermore, all the channel characteristics are dynamic, i.e. time variant, and they are specific to each pair of signal coupling points, or nodes, on each particular network.

In order to be successful, a broadband PLC technology must address these (and other) problems with their choice of frequency selection, modulation scheme(s), data packet structure and forward error correction (FEC).

The most interesting challenge, and one that forms the basis of this thesis, occurs as a result of the move to high frequency. Power distribution networks, optimised for the supply of electrical power at 50/60Hz, exhibit several non ideal radio frequency (RF) communication characteristics.

When a broadband HF signal is coupled onto a power distribution network, a fraction of the injected signal power will be radiated from that network into the surrounding environment.

This is an inevitable consequence of utilising spectrum within the HF band, aside from modifying the power distribution network, the only way of reducing radiated emissions is to reduce the transmitted power.

There are two main scenarios in which radiated emissions from PLC networks may cause interference to existing radio services.

Scenario 1 – Radio receiver system located in close proximity to a HF PLC network, will experience an increased radio noise floor and hence a reduced signal to noise ratio, assuming both systems to be operating within the same frequency bands.

Scenario 2 - where a large number of non-correlated PLC systems are installed in a given area, there will be an increased noise floor at a receiving site distant from all the PLC networks, again assuming an overlap of operating frequencies.

In both of these scenarios, the value of the noise increase, and therefore its significance to radio services, will depend on the PLC launch power spectral density (PSD), and the distance(s) between the PLC network(s) and the receiver / antenna location. Additionally, the noise increase will depend on the efficiency of the PLC network as a radiating structure. In scenario 1, near field electrical and magnetic coupling effects are likely to predominate. In scenario 2 only 'launched' electromagnetic energy and prevailing propagation conditions will be significant.

The noise floor increase caused by PLC is dependent on both the above parameters and the pre-existing noise floor, within the intended reception bandwidth. This increase could

potentially vary from a small fraction of a dB, (insignificant in radio terms), to many 10s of dBs.

Given existing power network infrastructures, and current PLC technology, the protection of co-located receivers / antennas, represented by scenario 1, can only realistically be achieved by careful frequency planning. Even where this is not enforced in legislation, it would be a matter of expediency for PLC designs to employ careful frequency selection. Thus, complaints from legitimate radio users in PLC deployment areas could be largely avoided.

In order to assess the risk to existing radio services in scenario 2, it has become necessary to investigate the radiating efficiency of power networks, and also to evaluate the signal strength regression in typical PLC deployment area topography.

1.12 Chapter 1 conclusions

Chapter 1 has introduced the field of power line communications and its development up to the present day. Modulation schemes have been evolved from simple examples to the more complex schemes typical of modern PLC systems.

Chapter 2 of this thesis will examine in more detail, the topic of Electromagnetic Compatibility with respect to the applicability of existing standards to the problems outlined above. Although these standards are not directly applicable to PLC networks, important principals can be drawn out and further developed later in the thesis.

Chapter 1

Chapter 2 also includes a brief discussion of antennas, with particular focus on there use in EMC measurement.

Chapter 3 undertakes a similar examination of the new EMC standards proposed in response to the first generation high frequency PLC systems.

2 Chapter 2 - Electromagnetic Compatibility

2.1 Introduction

Chapter 1 demonstrated that in order to provide power line communications systems capable of multi-megabit performance, it is necessary to use operating frequencies in the upper Medium Frequency (MF) and High Frequency (HF) bands – specifically between 1.6 and 30MHz. Figure 19 illustrates the frequency spectrum from 30kHz to 300 MHz.

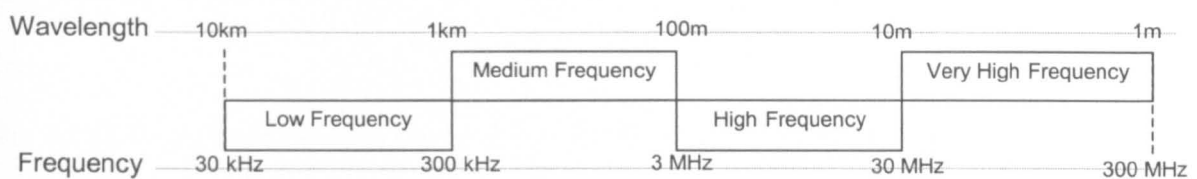


Figure 19 - The Electromagnetic Spectrum – LF to VHF

Some major technical issues are raised by this transition. Potentially most problematic of these issues is that of Electro-Magnetic Compatibility (EMC).

2.2 Electromagnetic Compatibility

Electromagnetic Compatibility can be defined as “The ability of a device, unit of equipment, or system to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbances to anything in that environment.” [21]

Clearly, these issues have been important since the early days of electronics and radio. Electrical and electronic devices, particularly radio receivers, have been in widespread use for many decades. Since there were surprisingly few serious problems, it might be concluded

that the early designs were relatively good with respect to EMC, or at least that commercial failure limited the production of poor designs.

In more recent years, there has been a rapid proliferation of electronic devices and systems, operating at increased frequencies and in a wider range of environments than ever before [21]. Combined with ever increasing reliance on electronics and telecommunication networks, this lead to the realisation that Electromagnetic Compatibility could no longer be left to chance.

EMC standards initially began to evolve in industries where reliable performance was critical to ensure safety, and where the environment could be readily defined and controlled. There is now a wide range of generalised EMC standards available, at both national and international levels. It is beyond the scope of this chapter to explain each one or even to list them all, but those that are relevant to PLC, even if not actually applicable are considered next.

Importantly, existing EMC standards deal with devices or systems with readily definable boundaries. With a PLC system, it is impossible to define in advance, all possible network configurations, and the extent of signal propagation within the network. After much interpretation of existing standards, it might be possible to demonstrate the EMC compliance, or otherwise, of individual items of PLC equipment. Unfortunately, this is only part of the overall picture when a PLC system as a whole is considered against the EMC definition quoted earlier in the chapter.

2.3 EMC Measurement Antennae

In order to more fully understand existing EMC testing techniques, and their potential applicability to PLC networks, it is necessary to examine some of the antenna types typically used in EMC compliance testing.

With conventional radio communications, it has always been preferable to design an antenna system for maximum sensitivity over the desired operating frequency range and maximum gain in the direction(s) of operation. At the very least the best balance of cost vs gain/sensitivity is the primary concern.

With EMC measurement antennae, a different set of design criteria is applied. In this case it is repeatability of measurement results that is of primary concern.

A flat frequency sensitivity across the widest possible frequency range is desired, along with a known quantifiable gain in the direction of measurement. Several antenna designs have been developed to meet these criteria.

2.3.1 *Magnetic Loop Antenna*

This type of antenna is typically used for testing up to 30 MHz, (wavelengths of 10 m and upwards) where antenna designs on conventional dipoles would be impractically large. This type of an antenna consists of a coil of wire, with a typical diameter of 1 meter. When excited in a magnetic field, the coil produces a voltage at the terminals proportional to the magnetic field.

The magnetic field strength H in amps / meter can be derived as follows :

$$H = \frac{4\pi \times 10^{-7} \times N \times A \times 2\pi F}{2.3.1.1} \quad \text{.....Equation 1}$$

Where E is the terminal voltage (volts), A is the area of the loop (m^2), F is the frequency of interest (Hz) and N is the number of turns.

As indicated in the above relationship, the sensitivity of the antenna varies with frequency, so it is common practice to use the antenna in conjunction with a powered pre-amplifier that provides the necessary compensation such that the frequency response becomes flat. This unit typically also provides impedance matching to 50Ω input impedance of EMC receiving equipment. [21]

By convention the magnetic field strength in amps / meter is converted to an electrical field strength, in volts / meter using the wave impedance of free space (377Ω). The validity of this assumption, for measurements typically made within the near field, is discussed later in this thesis.

Figure 20, shows a typical magnetic loop antenna setup. During the testing, the loop antenna can be rotated on its vertical axis for maximum field strength.

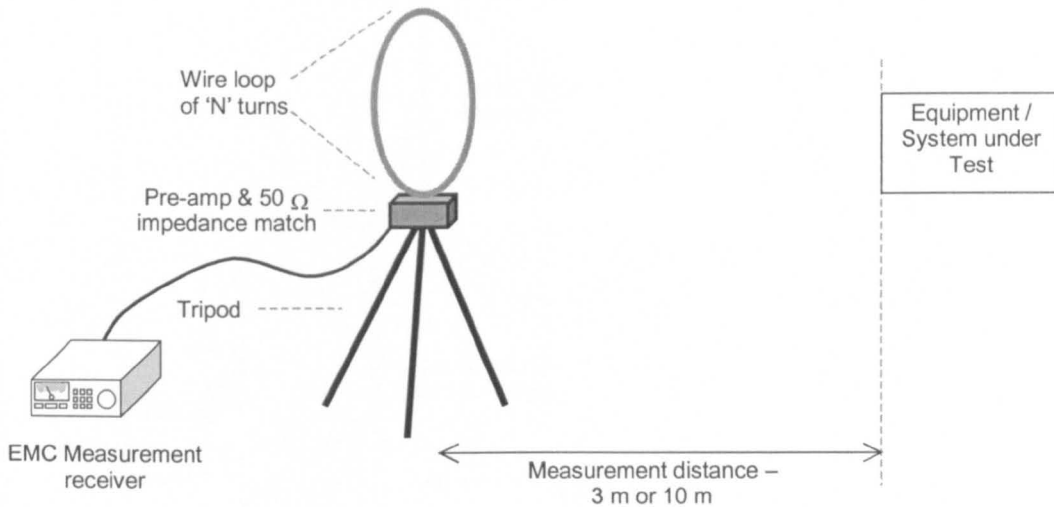


Figure 20 - Magnetic Loop Antenna measurement setup

At frequencies above 30 MHz (wavelengths below 10 m), dipoles or dipole based antenna designs are used. These can be used to directly measure electrical field strength. A conventional dipole can be tuned to a single frequency and used only at that frequency, or odd harmonics thereof. Broadband dipole antenna designs use pre-amplifier / impedance matching devices, similar to those used with magnetic loop antennas, to provide a flat frequency response across wide frequency ranges. Figure 21 shows a tuned dipole based measurement setup. [21]

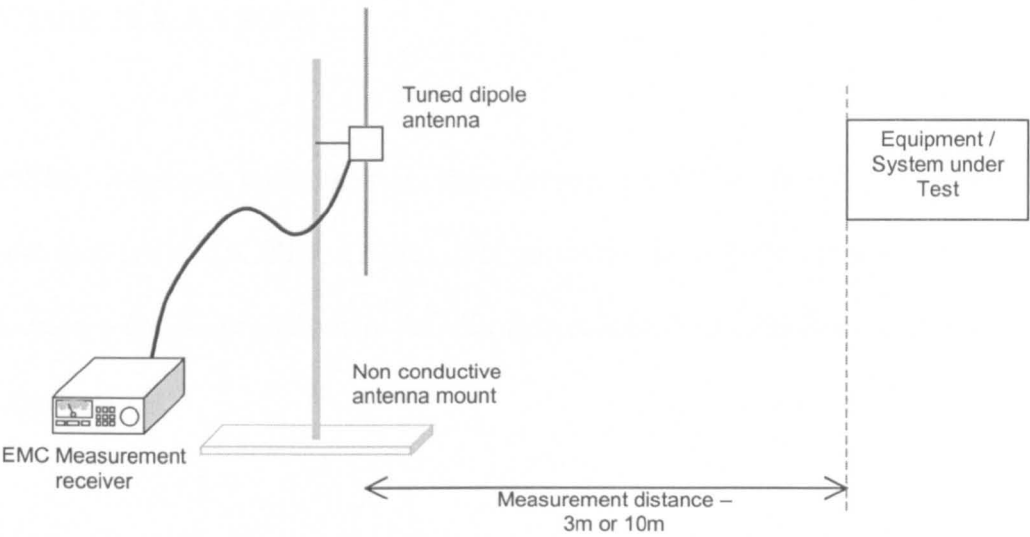


Figure 21 - EMC dipole antenna setup for discrete frequency measurements

2.4 CISPR 22 & EN 55022

Of the EMC standards available from the International Electro-technical Committee (IEC), the most applicable is CISPR 22 [22], last amended April 2003. This standard is entitled 'Information technology equipment - Radio disturbance characteristics - Limits and methods of measurement'.

According to the IEC, the intention of this standard is to 'establish uniform requirements for the radio disturbance level of the equipment contained in the scope, to fix limits of disturbance, to describe methods of measurement and to standardise operating conditions and interpretation of results'. [22]

It is clear from the definition of Information Technology Equipment (ITE) within CISPR 22 that telecommunications equipment falls within the scope of this standard. PLC systems are neither specifically excluded nor included but according to a strict interpretation of the ITE definition, would also fall within the scope of CISPR 22.

CISPR 22 classifies ITE equipment into two categories, according to its intended environment. Class B equipment is intended for a domestic environment, defined simply as an environment in which radio / television receivers are likely to be used within 10 m of the device or system in question. This would appear to be the most appropriate classification for any PLC system transmitting and/or receiving signals via the internal mains wiring of a residential or commercial building.

Class A devices are permitted significantly higher limits for conducted and radiated emissions but must display the following warning:

‘This is a class A product. In a domestic environment this product may cause radio interference in which case the user may be required to take adequate measures.’

For PLC systems operating over LV distribution networks, it may be felt acceptable to test to class A standard. In such cases, equipment installation could be strictly controlled. In addition, with this type of PLC system, signal penetration onto the building wiring may be limited - either through the application of filters, or due to the inherent attenuation caused by the electricity meter.

Class A & B Limits for conducted and radiated emissions, in a measurement bandwidth of 9 kHz, are shown in Table 1 and Table 2, below.

Frequency Range (MHz)	Limits (dBμV)	
	Quasi-peak	Average
0.15 to 0.5	79	66
0.5 to 30	73	60

Table 1 - Permitted conducted emissions on the mains port of class A equipment

Frequency Range (MHz)	Limits (dBμV)	
	Quasi-peak	Average
0.15 to 0.5	66 to 56 *	56 to 46 *
0.5 to 5	56	46
5 to 30	60	50
* Limit to decrease linearly with the logarithm of frequency		

Table 2 - Permitted conducted emissions on the mains port of class B equipment

Measurements are to be made using measurement receivers and Artificial Mains Networks (AMN) as defined in another IEC standard - CISPR 16 [23]. CISPR 22 requires the testing to be undertaken with the manufacturers intended method of connection to the LV supply. For PLC systems operating over LV distribution networks, it is not clear whether the second, ‘PLC signal only’, mains port would also be tested.

Although the above figures do provide limits, against which a PLC device could be tested, the limits are, by definition, at or below the existing noise floor. Thus, there would be enormous practical difficulties associated with designing a PLC system operating within these limits.

Radiated limits are only quoted for frequencies above 30MHz. This is perhaps an admission of the difficulties associated with accurately measuring field strength at such low frequencies.

The EN 55022 [24] standard from CENELEC, is a direct equivalent to CISPR22.

2.5 Electromagnetic compatibility of PLC devices or systems

The EMC standards discussed earlier in this chapter provide a useful insight into modern testing methods and limits. However, they are not directly applicable to PLC devices and systems.

Without an applicable standard, it is necessary for a PLC design to comply with EMCs principal objectives :

- Reliable operation within its intended electromagnetic environment
- Not inhibiting the reliable operation of other devices / systems within its electromagnetic environment

2.6 Defining the Equipment Under Test (EUT)

In order to consider the electromagnetic compatibility of a PLC design it is first necessary to define the 'Equipment Under Test' (EUT). For the majority of devices, the EUT can be defined with relative ease, (for a self contained, battery powered device it is obvious). For designs using mains power and with one or more communications ports, the EUT is generally defined as shown in Figure 22 and Figure 23 below.

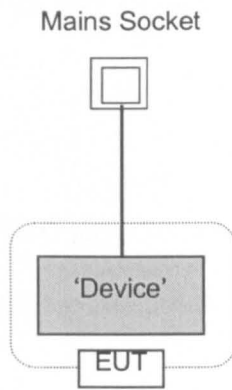


Figure 22 – Definition of EUT for devices without communication ports

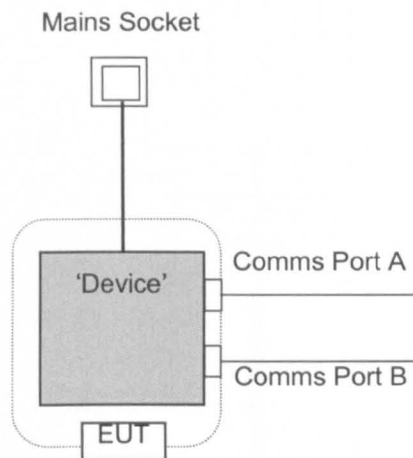


Figure 23 – Definition of EUT for devices with one or more communication ports

There are in fact two distinct types of PLC devices. The first type of PLC device connects directly to a standard mains socket from which it also draws electrical power. In terms of defining the EUT, it appears identical to the device shown in Figure 23, but in this case, the mains port is also used to couple data signals onto the power network. This presents a serious definition problem with the applicability of existing EMC standards. Since the port is directly connected to the mains network, it must be defined as a mains port. It is therefore subject to conducted emission limits across the whole frequency spectrum, including the band of frequencies in which the PLC system is operating. By definition therefore, the maximum

transmitted signal power, within any useable signal bandwidth, would have to be at or below the existing noise floor.

With the second type of device, the PLC signal is coupled onto the LV distribution network prior to the main fuse(s) and meter. Figure 24, defines the EUT for this type of device.

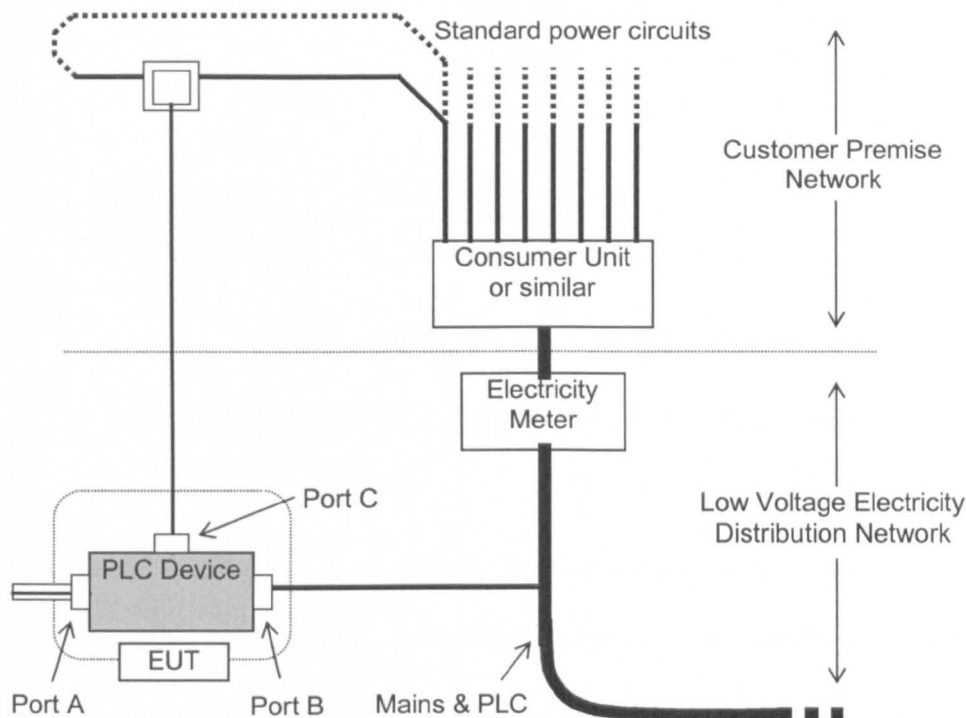


Figure 24 – EUT for PLC devices with separate power and PLC ports

This type of device is powered from a separate mains connection (Port C) that, along with telecomms port A, can comply with a conventional EMC standard, for example EN 55022. Port B clearly connects to an electricity power network, but rather than drawing energy from it, it transmits and receives modulated data signals. It therefore falls into neither definition.

This situation can be further complicated by the use of a separate coupling device, shown in Figure 25.

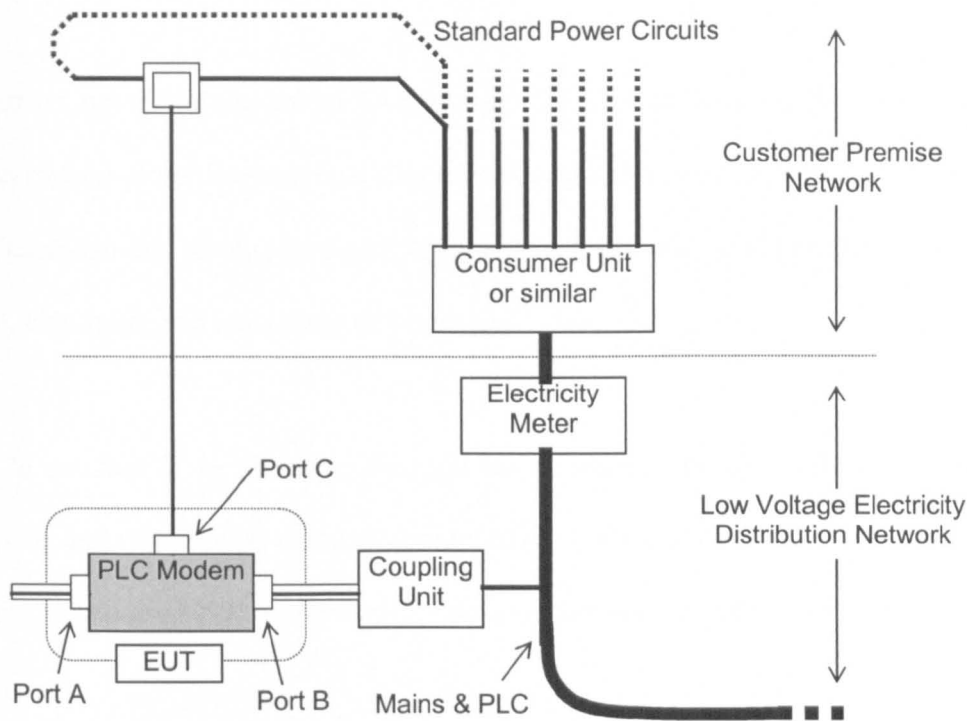


Figure 25 – EUT for PLC device with separate coupling unit

In this system, the Coupling unit is an entirely passive device, drawing no energy from either the electricity distribution network, or the PLC modem. It simply couples RF signals between Port B on the PLC modem, now clearly definable as a telecoms port, and the electricity distribution network. This arrangement would allow the PLC modem, taken in isolation, to be compliance tested against any EMC standard deemed necessary. However, this solution fails to address the Electromagnetic Compatibility of the PLC system as a whole, including the mains network. Work is now underway within the standards bodies to resolve these issues and provide a series of new standards, with which PLC and similar systems can claim compliance.

2.7 The International Electrotechnical Commission (IEC)

The International Electrotechnical Commission (IEC) was founded in 1906, following a resolution passed at the International Electrical Congress held in St. Louis (USA) in 1904. It seeks to establish and promulgate international standards and good practice in all aspects of electrical, electronic and associated technologies.

The IEC's mission is to promote, through its members, international co-operation on all questions of electrotechnical standardization and related matters, such as the assessment of conformity to standards, in the fields of electricity, electronics and related technologies. [25]

The IEC charter embraces all electrotechnologies including electronics, magnetics and electromagnetics, electroacoustics, multimedia, telecommunication, and energy production and distribution, as well as associated general disciplines such as terminology and symbols, electromagnetic compatibility, measurement and performance, dependability, design and development, safety and the environment. [25]

Figure 26, shows the structure of international standard creation within the IEC.

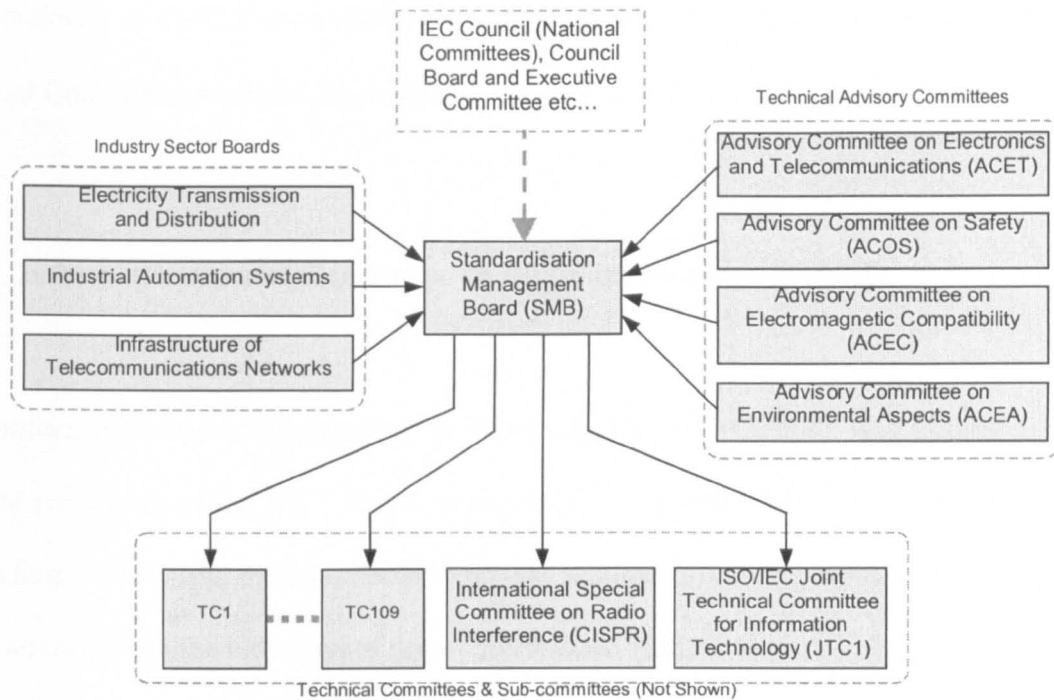


Figure 26 - Standards creation structure within the IEC

Within the IEC structure, low frequency Power Line Communications (PLC) for power network automation, has historically been considered by Technical Committee 57, Power System Control and Associated Communications. A number of relevant standards have been published, most notably IEC61334. [26]

Technical Committee 77, Electromagnetic Compatibility considers, in generic terms, conducted immunity, and conducted emissions below 9kHz for example mains harmonics and voltage fluctuations. TC77 also considers, in co-operation with CISPR, conducted emissions above 9kHz that are not covered by the CISPR [10] standard, including mains signalling (up to 148.5kHz only). Additionally, this committee considers the safety aspects of EMC, and has a horizontal function in providing this information to other technical committees. [27]

The majority of EMC standardisation work within the IEC is carried out by the Information Special Committee on Radio Interference (CISPR), and its various sub-committees.

2.8 Information Special Committee on Radio Interference (CISPR)

The Information Special Committee on Radio Interference, (CISPR), was formed in 1934 as a joint committee of the IEC. It is now a special committee of the IEC, with a membership including not just IEC national committees but also the following international organisations with an interest in the reduction of radio interference. [28]

- European Broadcasting Union (EBU)
- International Amateur Radio Union (IARU)
- International Conference on Large Electric Systems (CIGRE)
- International Union of Electro-heat (UIE)
- International Union of Producers and Distributors of Electrical Power (EURELECTRIC)
- International Union of Public Transport (UITP)
- International Union of Railways (UIC)

CISPR also collaborates directly with the International Telecommunication Union (ITU), the International Civil Organisation (ICAO) and the European Telecommunication Standards Institute (ETSI). Within the IEC, CISPR liaises closely with TC77, under the co-ordination of the Advisory Committee on Electromagnetic Compatibility (ACEC).

Chapter 2

CISPR stated primary aim is the protection of radio services operating in the frequency range 9kHz to 400GHz. By brokering international agreement on radiated emission limits for all types of unintentional radiators, and on immunity limits for susceptible devices, CISPR facilitates international trade. CISPR currently has a total of 7 sub-committees and 15 working groups, the scope of which are explained in IEC publication CISPR 10. The CISPR sub-committees are [26]:

- CISPR/SC A – Radio interference measurements and statistical methods.
- CISPR/SC B – Interference relating to industrial, scientific and medical radio-frequency apparatus.
- CISPR/SC C – Interference relating to overhead power lines, high voltage equipment and electric traction systems.
- CISPR/SC D – Interference relating to motor vehicles and internal combustion engines.
- CISPR/SC F – Interference relating to household appliances, tools, lighting equipment and similar apparatus.
- CISPR/SC H – Limits for the protection of radio services.
- CISPR/SC I – Electromagnetic compatibility of information technology equipment, multimedia equipment and receivers. (CISPR/SC E and CISPR/SC G are now combined in this sub-committee)
- CISPR/SC S – Steering committee of CISPR
CISPR Standards relevant to PLC

Of particular interest to the manufacturers and operators of PLC equipment is CISPR/1 and, in particular, working groups (WG) 3 and 4 of this sub-committee dealing with emission from, and immunity of information technology equipment (ITE) respectively.

By defining a PLC modem as an information technology product, and accepting the applicability of the CISPR22 (ITE emissions limits) and CISPR24 (ITE immunity limits), it is possible to demonstrate compliance or otherwise with an internationally recognised standards. The same logic can be applied to the equivalent CENELEC standards EN55022 and EN55024, which are applicable throughout the European Union.

CISPR22 and EN55022 specify limits and measurement methods for conducted emission from mains and telecommunication port(s) of the equipment under test (EUT), within the frequency range 9kHz to 30MHz. For the mains port however, no distinction is made between undesired conducted emissions, and desired PLC signals. Thus for a PLC modem, transmitting high frequency signals via a combined mains and telecommunication port within CISPR22 specified limits, might preclude reliable communication.

2.9 European Committee for Electrotechnical Standardization (CENELEC)

The European Committee for Electrotechnical Standardization, (CENELEC), was established in 1973 as a non-profit making body organisation under Belgian law. Directive 83/189/EEC from the European Commission recognises CENELEC as the principal European Standards Organisation within this field. [29]

Many CENELEC standards originate at an international level; indeed its relationship with the IEC has been discussed earlier in this thesis. In fact, at the end of the year 2000, 66% of all published CENELEC standards were direct equivalents of IEC standards, and a further 8% were based on IEC work and results. [30]

The structure of standardisation activities within CENELEC is similar to that of the IEC. A Technical Board (BT) co-ordinates the work of the technical bodies. Technical Committees (TCs) are established, are given titles and scopes, and are dissolved by the BT as required. TCs themselves may, after approval from the BT, set up further Sub-Committees (SCs). The BT is also able to establish Technical Board Task Forces (BTTFs) and Technical Board Working Groups (BTWGs), in order to carry out specific short-term tasks with defined target dates.

There are four potential starting points for a harmonised CENELEC standard [27]:

- An initial document comes from the International Electrotechnical Commission (80% of cases).
- A document of European origin arises in one of CENELEC's own technical bodies.
- A first draft of a European document comes from one of CENELEC's co-operating partners.
- The National Committees themselves. Under the 'Vilamoura Procedure', the National Committees (NCs) have agreed to notify CENELEC when they are planning any new work. CENELEC can, if it wants, take on this work.

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After a 6-month consultation process, during which NCs have the opportunity to comment on the document, a final draft is prepared and sent for voting.

Within CENELEC, sub-committee (SC) 205A deals with mains signalling systems. Until recently, this committee has discussed low frequency systems, operating between 3kHz and 148.5kHz, and the EN50065 suite of standards is close to completion. The scope of this technical committee has now been extended to cover high frequency. The new scope [29] reads :

“To prepare harmonized standards for communication systems using low voltage electricity supply lines or the wiring of buildings as a transmission medium and using frequencies above 3 kHz and up to 30 MHz. This includes the allocation of frequency bands for signal transmissions on the mains.”

Working group 10 of SC 205A has now been set up to address the additional technical issues associated with the above increase in scope.

CENELEC standard EN55022 specifies conducted emission limits for information technology equipment, the problems posed by its application of its direct IEC equivalent, CISPR22, to PLC have already been discussed.

2.10 The European Telecommunications Standards Institute (ETSI)

The European Telecommunications Standards Institute, (ETSI), is a not for profit organisation whose mission is to produce the telecommunications standards that will be used for decades to come throughout Europe and the world [31]. It was created in 1998 by the European Conference of Postal and Telecommunications Administrations, (CEPT), as a home for CEPTs ongoing telecommunication standardisation activity.

ETSI currently has 874 members, spread over 54 European and non-European countries, and comprising administrations, network operators, manufacturers, service providers, research bodies and users. Standardisation activity is undertaken by, and on behalf of ETSIs membership, ensuring that it maintains a close alignment with market needs. ETSI Membership, including full voting rights in technical bodies, is open to any company or organisation with an interest in the creation of telecommunications standards.

An ETSI project entitled “Powerline Telecommunications” has been established with the following scope: [32]

“The project will progress the necessary standards and specifications to cover the provision of voice and data services over the mains power transmission and distribution network and/or in-building electricity wiring. The standards will be developed in sufficient detail to allow interoperability between equipment from different manufacturers and co-existence of multiple powerline systems within the same environment. Harmonised Standards will be developed to allow presumption of conformity with the relevant EU/EC Directives.”

ETSI PLT is a wide ranging project, covering such diverse subject areas as service requirements, interfaces, reference configurations, protocols and media access, physical layer characteristics, spectrum management, encryption, conformity and safety. In fact, the majority of its work has been shared with other technical bodies. For example, spectrum management is being discussed within the spectrum engineering working group 35 (SE35) of the European Radiocommunications Committee (ERC), a part of ETSI's parent organisation - CEPT. EMC issues fall within the remit of a newly formed ETSI/CENELEC joint working group, and protocol issues have been divided between CENELEC SC205A, (lower layers), and ETSI PLT, (higher layers).

2.11 Chapter 2 conclusions

This chapter has introduced the topic of electromagnetic compatibility, and detailed the principals that underlie all EMC regulation.

This chapter also examined existing international standards covering EMC. In one case, with FCC part 15, clauses are in place to cover high frequency power line communications. In all other cases high frequency power line communication is not discussed at all.

Importantly, none of the examined standards anticipated widespread deployment of PLT systems in domestic and commercial environments. In order to create EMC standards that are specific to PLT, or to extend existing EMC standards to cover PLT systems, it becomes critical to understand the relationships between injected signal power, or power spectral density within designated operating frequency bands, and near / far field radiated emissions. Once this relationship is fully quantified, it will be possible to create EMC standards that can mitigate interference to existing radio systems and electronic devices to the extent that problems can be avoided.

This chapter has also given further insight into the structure of the world's standards organisations and the processes through which new standards are produced. It is through these bodies that future EMC standards are being created.

Chapter 3 will examine the new EMC standards that are being developed in the above standards bodies to apply to PLT and similar networks where retro-fit of HF modem equipment onto existing (non-ideal) networks is proposed, for example VDSL.

3 Chapter 3 - Proposed PLC standards

In the previous chapter, a number of existing EMC standards were discussed, along with their varying degrees of applicability to HF PLT systems. It was demonstrated that, in their current form, none were completely suitable for the control and regulation of radiated and conducted emissions from PLT systems operating in the HF band – 1.6 to 30 MHz. As discussed, European standardisation activity is now being driven forward by CENELEC SC205A and ETSI PLT. Work is also continuing at an international level within IEC CISPR 22.

3.1 Drivers for new international standards

The lack of EMC standards covering PLC systems had been recognised in Europe and the rest of the world by the late 1990s. At this time, trials of HF PLT equipment were taking place using first generation equipment from a number of companies world wide, most notably in Europe, Nor.Web DPL Ltd. This company was formed as a joint venture between Canadian based 'Nortel Networks' and British utility company 'United Utilities'. Their product, capable of a shared 1Mbit/s data rate, was undergoing extensive trials on sites in Germany, Sweden and the UK. Initial trials proved a technical success and no significant interference issues arose. Two further companies, Siemens and Ascom were also looking to trial PLT equipment, and there was pressure from all involved to allow full commercial deployment as soon as practical.

Understandably, this raised concerns within the international radio community. Wide spread deployment of HF PLT, along with similar technologies such as VDSL, was seen as a potential new source of interference to long distance HF wireless communications.

Both technologies proposed to use signals in the HF band on existing wiring systems that were not originally installed and developed to support such signals. In the case of VDSL, some protection from radiated emissions was offered by the fact that the wiring was of 'twisted pair' construction. With PLC, the pseudo-coaxial nature of the LV cables would provide some degree of shielding. These effects, however, were not easily quantifiable.

Proposed standards emerged from the Radiocommunications Agency (RA), now part of OFCOM, in the UK and from RegTP in Germany.

3.2 Proposed radiated emission limits and measurement techniques applied to PLC

3.2.1 MPT1570

At UK national level, the Radio Communications Agency proposed a new standard, MPT1570. This standard specified radiated emission limits, to be measured in 9kHz bandwidth, from wired telecommunication networks (a definition that includes PLT, VDSL and all similar technologies). MPT1570 was intended as a mitigation standard in case of an actual interference incident, rather than used as an 'equipment compliance' standard. Measurement technique and equipment is fully specified, with reference to existing EMC standard - CISPR 16-1. The levels specified by MPT1570 are summarised in Table 3, below:

Standard	Frequency Band	Radiated Emission Limit (dB μ A/m)	Radiated Emission Limit (μ V/m) *	Measured At (m)
MPT1570 (Jan 2003)	150kHz to 1.6MHz	-11.5-40*log f(MHz) (peak)	40-7.7*log f (MHz) (peak)	3
MPT1570 (Draft of Feb 2000)	1.6MHz to 30MHz	20-7.7*log f(MHz) (peak)	-31.5-7.7*log f (MHz) (peak)	3

Table 3 – Proposed MPT1570 Emission limits

* Although the actual limit is specified in dB μ A/m, the equivalent electrical field strength, using a conversion factor of 51.5dB, derived from the wave impedance of free space, is also shown above. This is shown for the purposes of direct comparison with other standards.

In its current form there is no specified limit above 1.6 MHz, but the limit initially proposed in a draft of Feb 2000 is shown in Table 3, above, for the purposes of comparison with contemporary standards.

3.2.2 Usage Provision 30 (NB30)

The Regulatory Authority for Telecommunications and Posts (RegTP) in Germany proposed Usage Provision 30 (NB30), which has since been approved by the German Parliament. It is not technically a standard, but it does give a blanket approval for deployment, within Germany, of PLC systems that meet two conditions. These conditions are:

First - that the system does not make use of frequencies used for the operation of safety related radio communications.

Second - that at all points along the network, the interfering field strength does not exceed $40 - 8.8 \cdot \log(\text{Frequency in MHz})$ dB μ V/m measured in a 9kHz bandwidth at 3m. Measurement is to be carried out as specified in RegTP standard 322 MV 05.

If either of the above conditions cannot be met, then deployment can only be approved on a site by site basis.

There has been much debate over the past few years about the definition of safety related telecommunications, and the exact frequency bands captured by this statement.

3.2.3 Comparison of MPT1570 and NB30 with the FCC Part 15

The existing FCC part 15 standard [33], which is applicable throughout the North America market does specify radiated emission limits between 1MHz and 30 MHz, albeit at a different measurement distance to the 3m specified in MPT1570 and NB30.

A comparison of the three standards on the same chart of permitted field strength against frequency illustrates the wide variation in limits, each claimed to be the minimum necessary to ensure an acceptably small level of interference to existing radio users. The differences between Europe and North America are attributable to a combination of differing geography. For example, the population densities tend to be much higher across most of Europe. There is also a differing regulatory approach to such problems. The FCC limits were specified a number of years ago when deployment of such systems was likely to be on an extremely limited basis, and the potential for interference therefore was small.

The FCC specifies a measurement distance of 30 meters. With the limited deployments considered by the standard, it would be much easier to do field strength measurements at this distance, without the risk of inaccuracies caused by adjacent PLC networks. Figure 27, below compares the MPT1570, NB30 and FCC part 15 emission limits.

In the case of FCC Part 15, the limit was extrapolated from a 30m to a 3m measurement distance using a ‘worst case’ factor of 20dB per decade. This is permitted under the standard itself, indeed it may be possible to use a larger regression factor if it can be demonstrated that the figure is representative.

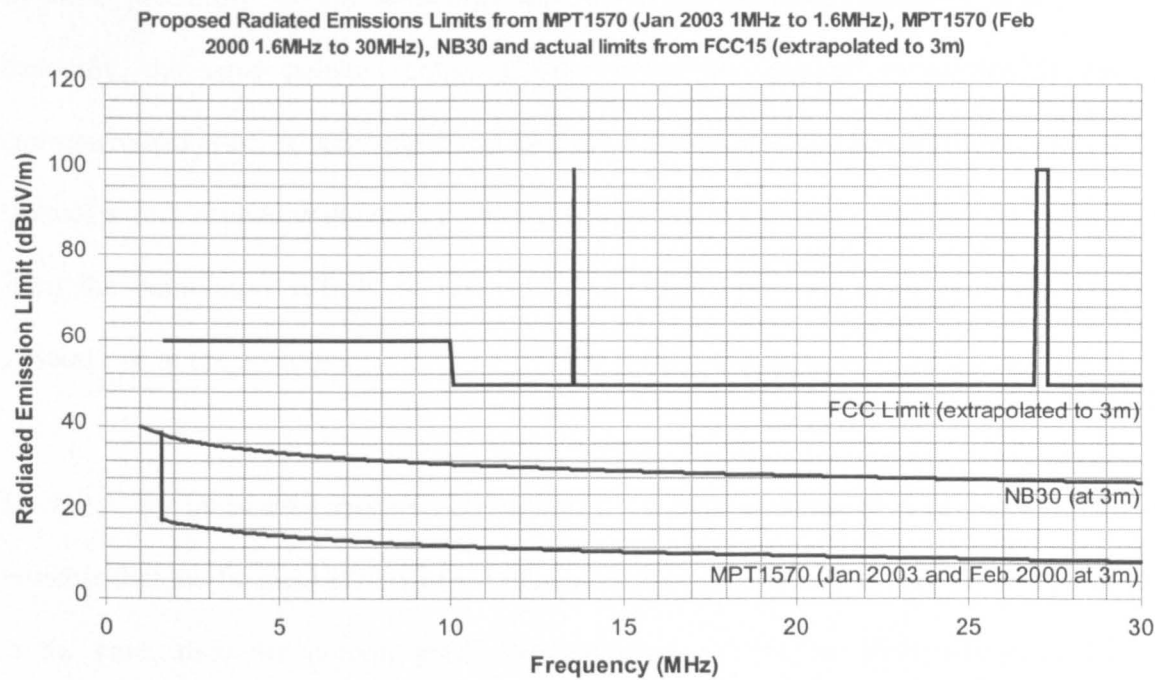


Figure 27 - Radiated emission limits for PLC - MPT1570, NB30 and FCC 15

MPT1570 and NB30 were designed to protect licensed users of the radio spectrum. In setting the precise emission limits, and measurement distances, many factors need to be considered.

By convention, measurement distances are typically set at 3m, 10m or 30m which ever is most appropriate to the frequency band and measurement method. Unfortunately, the persuasiveness of power line networks may require the measurement distance for these standards to be set at 3m despite the frequency band involved. Only at this short distance, can the effects of signal addition and cancellation caused by the relative proximity of other network elements be minimised.

Taking the measurement distance to be 3m, the radiated emission limit would fulfil two main purposes. Firstly, it would offer protection to HF receiving equipment and antennas located in close proximity to the radiating network, i.e. within the inductive near field [34]. Secondly, the same radiated emission limit would be required to protect HF receiving equipment and attached antennas from far field electromagnetic radiation, received via direct (line of sight), ground wave or sky wave propagation. In this case, the problem would come from the cumulative effects of many PLC systems operating simultaneously, causing a general rise in the noise floor.

However, by measuring radiated emissions from PLC systems at a distance of 3m, it is possible that the far field effects of HF PLC systems are not being accurately assessed. If this is the case, then the current predictive modelling of the far field effects of HF PLC deployment may also be inaccurate.

3.3 High Frequency radio propagation modes

As discussed above, practical considerations will require the radiated emission limits to be specified at a relatively short measurement distance – probably within the near field. Within the context of HF PLC systems, the $\lambda/2\pi$ boundary will be between 2 and 48 meters.

It is necessary to specify a radiated emission limit, in the above circumstances, that adequately protects remote radio sites against cumulative interference from multiple PLC networks. This requires a more complete understanding of the radio propagation modes within the HF band. Within the 1.6 to 30 MHz frequency band, three propagation modes – ground wave, space wave and sky wave pre-dominate. [35] These propagation modes are illustrated in Figure 28 below.

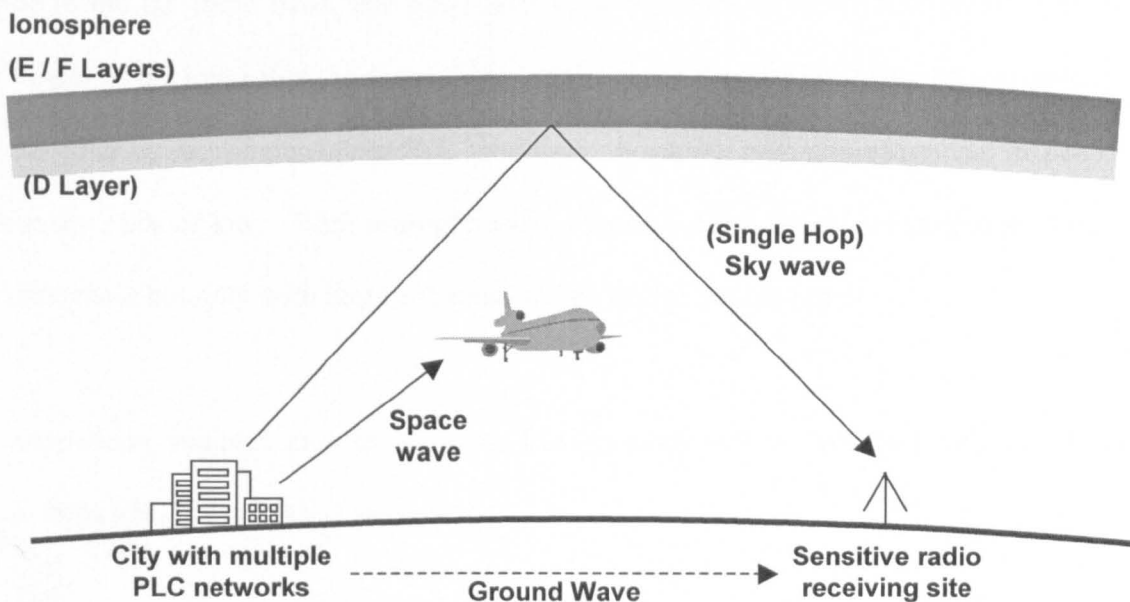


Figure 28 - High Frequency Propagation modes

In the above example, Ground wave propagation will only occur at frequencies up to approx 10 MHz. It is also dependant on the geology between the transmitter and receiver. Space wave is mathematically predictable if the antenna (k) factor of the PLC system in the direction of the receiving antenna is known or can be approximated. As a HF propagation mode though, its effectiveness between ground based sites is limited to only a few km by the curvature of the earth.

In the context of potential HF PLC interference to distant receiving sites, single hop sky wave propagation would appear to be the dominant propagation mode.

With this propagation mode, the earth's ionosphere creates a reflective (and partially refractive) boundary. Thus, radio signals transmitted from the earth's surface towards outer space are re-directed back down to another point on the earth's surface. This phenomena, unique to the HF radio band, has been used since the early days of radio communications. The height of the ionosphere is dependant on the time of day and its thickness varies with the 11 year solar cycle. Under favourable conditions, however, communication can be achieved over many 100s of km. With multiple earth – space – earth 'hops', even greater distances are achievable but only with large antennas and powerful transmitters.

By comparison, radiated interference from PLC systems will be launched with very limited power from a highly non-ideal antenna.

3.4 European Community Mandate M-313

The MPT1570 and NB30 standards were welcomed by many on both sides of the debate, as they did at least specify actual field strength limits and measurement methods. The debate was now able to move onto discussion of whether the new limits were too low – inhibiting the commercial viability of PLT, or too high – risking widespread interference to radio services.

There was a desire within the UK government and indeed internationally to see PLT become a market rival to the xDSL technologies which were largely under the control of incumbent telecomms operators.

Neither side of the debate was entirely satisfied with the standards in their current form, and it was generally accepted that a single harmonised European standard would be required in any case.

In August 2001, the European Community (EC) issued mandate M-313 to CENELEC and ETSI, requiring the development of a new EMC standard specifically covering telecommunication networks, including power lines, co-axial cable networks and telephone networks.

This standard would specify limits and testing methodology with sufficient detail such that compliance with the overall EMC directive (89/336/EEC) could be assured. It also required the withdrawal of conflicting national standards.

Interestingly, it is possible that MPT1570 and NB30 could remain in force in the UK and Germany respectively, since neither has the status of a full national standard. It is thought that the latest generations of PLC equipment may incorporate some dynamic flexibility in their choice of operating frequency(s). Thus they could comply with both the new European standard resulting from M-313 and, where specific interference issues arise, the applicable national standard.

Although still at the draft stage, a limit of 55.5 dB μ V/m at 3 m (peak in 9 kHz) has been proposed across the 1.6 MHz to 30 MHz band. This is an extrapolation of a CISPR 22 derived figure of 22.5 dB μ V/m at 30 m, using an extrapolation factor of 33 dB per decade.

The draft also specifies that measurements are to be undertaken at a minimum of three different open field sites that are representative of a typical installation. In all cases, the axial of maximum field strength from the network, in other words, the direction in which the field strength is highest at 3 m, must be established and used.

3.5 Benchmarking HF PLC networks in the far field

From the examination of existing and proposed EMC standards, it is clear that a detailed measurement program evaluating far field radiated emissions, both direct and ground wave propagation modes, from HF PLC systems is required. This presents a number of technical difficulties, some of which have already been discussed [36], others are discussed below.

In order to assess far field radiated emission profiles from HF PLC networks in general, rather than from a particular system, there are a number of experimental parameters that would need to be defined. This is done throughout the remainder of chapter 3.

3.5.1 *Test signal modulation, frequency and power level*

The measurement program would need to take place outside of the relatively noise free EMC laboratory environment and the receiving equipment will need to be located at some considerable distance from the radiating network in order to be sure that true 'far field' radiated interference is being measured. There is a requirement not to cause interference to licensed HF radio users, particularly emergency services. The experimental program must also be able to cope with both an increased noise floor and with 'interfering' HF transmissions from broadcast and amateur radio services.

By adopting a simple continuous carrier wave test signal, two main advantages would be realised. First, the test signal would contain all its energy within only a few Hz. By careful frequency selection, the interference to existing HF users could be minimised. Secondly, the

required receiver bandwidth can be very much lower than the 9 kHz typically used for HF voice transmission, allowing the receiver to more easily avoid external interfering signals, and to provide an apparent reduction in the noise floor.

Even if a suitable wideband receiving antenna was available, it would be extremely difficult to use a 'swept frequency' approach to benchmarking far field HF PLC emissions, because of the interference this would cause to other HF users. A good compromise would be to select a number of discrete frequencies, covering the band 1 MHz to 30 MHz, at which the test signal can safely be injected. Amateur radio bands, (LF band edges), might be particularly suitable for this purpose, providing available frequencies, distributed about the HF spectrum, at approximately 1.8, 3.5, 7, 10, 14, 18, 21, 25 and 28 MHz [37]

For any given frequency, the received signal power will be directly proportional to the injected signal power. In order to study the far field emission, over a distance of a few hundred meters, it would be necessary to use an injected signal with a power spectral density considerably higher than that of the proposed HF PLC systems. From practical experience, injected power levels in the range 0 dBm to +20 dBm, with PSDs -40 dBm/Hz and -20 dBm/Hz (in 9 kHz) respectively, would seem to be the most suitable. Above +20 dBm, (an actual injected signal power of 0.1 W), the presence of a licensed radio operator would be required.

Some HF PLC systems couple signals onto the mains in a live to earth configuration, others use a live to neutral arrangement. With LV distribution networks, phase to phase signal coupling is also possible. It is difficult to predict which configuration, if any, will cause the

highest level of radiated emission. For a comprehensive assessment of radiated emissions, all configurations should be tested on a number of different networks.

3.5.2 Signal injection into the mains network

In order to inject the signal onto the mains network, a coupling unit is required. This device must provide complete 50 / 60 Hz isolation between the phase and neutral conductors and the signal generating equipment and operator, but at the same time provide low signal attenuation at HF.

The selection of signal injection locations for an 'in building' network is relatively easy - any accessible socket on the network should be suitable. For LV distribution systems, injection points close to the LV transformer, and the meter position should be representative of proposed HF PLC systems.

At high frequency, the impedance of any given point on an 'in building' mains network has an average impedance modulus of close to 50 Ω , with LV distribution networks typically exhibiting a lower average. However, there is a wide impedance range, and a significant impedance mismatch at the signal injection point is possible. One approach to this problem is to design the coupling unit such that it can be 'tuned' to the impedance of the signal injection point. In this case, the mismatch is eliminated and all the signal power can be injected into the mains network. Alternatively, if the mismatch can be quantified at each signal injection point and frequency, its effect can be considered in the final calculation of radiated emission.

3.5.3 *Radiated emission measurement antenna*

Active loop antennas and active rod antennas with ground planes, although useful for near field measurements in a controlled environment, are not ideal in the far field. As an alternative, a narrowband HF antenna of traditional design is proposed [38] [39].

An inverted 'V' dipole antenna is constructed from a single mast supporting a half wave dipole. The ends of the dipole are attached via suitable insulators to convenient anchor points on the ground. A HF 1:1 balun is used to connect the unbalanced 50 Ω feedline to the dipole terminals. The length of each half of the dipole, initially set to 1/4 of the wavelength, is subsequently reduced to bring the dipole into resonance at the desired frequency. In this state, the antenna has an impedance of approximately 50 Ω .

The usefulness of this antenna design increases when harmonic effects are considered. In addition to its fundamental frequency, the dipole will also be resonant, i.e. suitable for use, at 3rd, 5th 7th etc. harmonics. Thus an antenna designed for 3.5 MHz, will also be usable at 10.5, 17.5 and 24.5 MHz.

A second dipole, attached to the same balun terminals as the first, and supported by the same mast can be independently tuned to another resonant frequency at an even harmonic of the first dipole. At each dipole's resonant frequency, and its odd harmonics, the other dipole will present a high impedance to the balun. Thus an antenna could be constructed, using a single balun and feedline, and offering acceptable performance at 3.5, 7, 10.5, 17.5, 21 and 24.5 MHz. Other frequency combinations would be just as easy to realise. This arrangement is shown in Figure 29, below.

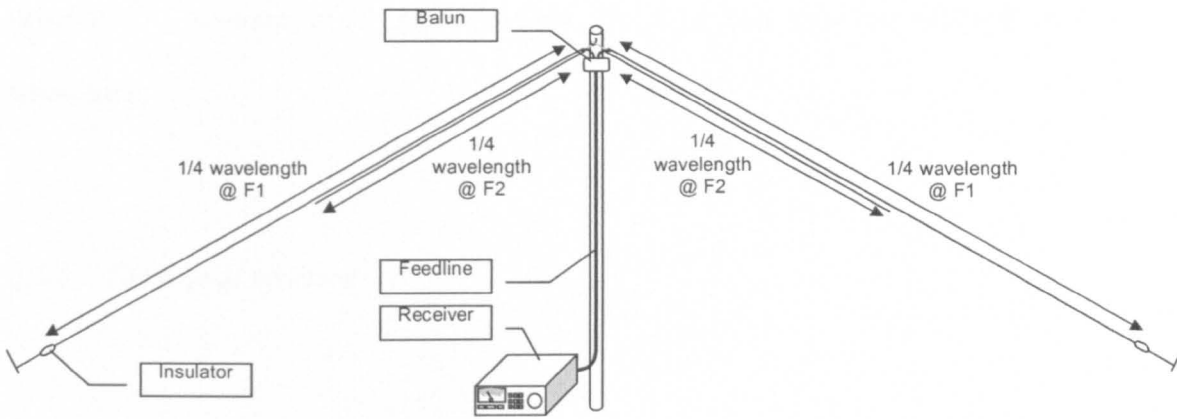


Figure 29 - The proposed multi-band inverted 'V' dipole antenna

This configuration is primarily horizontally polarised, and for accurate measurement should be orientated perpendicular to the direction of the radiating network. The antenna is of modest cost, and could be erected and dismantled in only a few minutes.

A further alternative antenna, offering complete portability would be one of the commercially available 'loaded' vertical whip antenna, mounted on the roof of a vehicle. Such antennas are usually helically wound, in order to reduce the physical length to 2 m or less. This vertically polarised arrangement is particularly suitable for examining HF PLC ground wave propagation, and the antenna might be calibrated where absolute field strength values are required.

3.5.4 The Measurement distance

With distances of up to 300 meters, the maximum practical range over which HF signal regression from this experiment can be measured, the most appropriate way to determine measurement distance would be to use a GPS navigation system. By recording the positions of the signal injection point, and the measurement point, the distance between them could be

calculated. Accuracy's of approximately +/- 5 m can now be realised with standard equipment.

3.5.5 *The Signal receiver*

Calibrated spectrum analysers and/or receivers are the standard tool for EMC measurement. With a spectrum analyser, the signal level and composition, across an entire frequency range, may be observed within a user selectable receiving bandwidth, usually 9 kHz. In this application however, the cost and non portability of such equipment is a serious disadvantage.

The alternative approach, given the need to measure signal strength only at discrete frequencies, is to use a portable HF receiver with a signal strength meter, tuned to the frequency of interest. In order to convert the signal strength meter readings to accurate signal power values, it is necessary to 'calibrate' the receiver. This process does not truly calibrate the receiver's signal strength meter, since no actual adjustment is made to the receiver itself. A calibrated signal generator should be connected to the antenna input of the receiver. Taking care not to overload the sensitive receiver input stage, the signal generator output power should be adjusted such that the exact signal power from the signal generator, for each signal strength reading on the receiver, S1, S2, S3 can be measured and recorded. Since the performance of the receiver can vary significantly with frequency, the process must be carried out at each frequency of interest. The end result of this process is a series of S meter to dBm conversion charts, each valid at a single frequency.

3.5.6 Calculation of field strength

With the received signal power now known, at each measurement position and frequency, and assuming a good antenna / feedline / receiver impedance match, the field strength can be calculated by examining the following relationships. The received signal power in terms of the effective capture area (A_e) of the receiving antenna in m^2 , and the power flux density (P_d) of the received signal in $watts/m^2$ is given by $P_R = A_e P_d$

Equation 2, below. [40]

$$P_R = A_e P_d \text{ Equation 2}$$

Where the numerical receiving antenna gain (G_R) in the direction of the signal source, is known, for example a half wave dipole has a numerical gain of 1.64, the effective capture area can be calculated as follows [38] [40] :

$$A_e = \frac{G_R \lambda^2}{4\pi} \text{ Equation 3}$$

Otherwise, the effective capture area could be established by measurement. The power flux density (P_d) at a point with known electromagnetic field strength can be determined from the following equation :

$$P_d = \frac{E^2}{120\pi} \text{ Equation 4}$$

In the above equation, E is the electromagnetic field strength in V/m, and 120π represents the impedance of free space. By inserting into Equation 2, and re-arrangement the following useful relationships of electric field strength may be defined :

$$E = \sqrt{\frac{120\pi P_R}{A_e}} \dots\dots\dots \text{Equation 5}$$

$$E = \frac{\pi}{\lambda} \sqrt{\frac{480P_R}{G_R}} \dots\dots\dots \text{Equation 6}$$

Equation 5 can be used to calculate the electrical field strength in V/m, where the wavelength, effective capture area of the receiving antenna gain and received signal power in watts are known. Equation 6 can be used when the receiving antenna's gain, in the direction of the signal source, is known.

With a practical measurement system, the signal power measured at the receiver, could be significantly affected by signal losses. There are likely to be impedance mismatches within the antenna/feedline/receiver signal path. There could also be signal loss within the feedline and, where present, within the HF balun. From experience, these factors can account for up to 3dB of signal loss within the system. A more accurate figure for received signal power, and therefore field strength, could be found by introducing a factor of 0.5 into Equation 1. Alternatively, where the exact signal loss within the system can be quantified, an exact correction factor can be used.

3.6 Chapter 3 conclusions

In the first part of this chapter, two developing EMC standards, MPT1570 and NB30 were examined. Both of these standards were proposed in response to understandable concern from the radio community, that trial PLT systems and even large scale PLT deployment might take place before any interference protection measures could be taken. The adequacy of the protection offered by these standards was the subject of considerable debate within the standards bodies concerned, but at least they provided a basis for measuring and limiting interference from the early PLT trials in the UK and Germany.

When all HF radio propagation modes are examined, it becomes obvious that local (near field) radio interference, as measured by the above standards, is not the only potential EMC problem. Direct, ground wave and sky wave HF propagation, over larger distances might cause an aggregation of interference from numerous PLT networks to increase the noise floor at a distant point, to an extent which would be considered interference. It is therefore necessary to quantify both the effectiveness of typical mains networks as transmitting antennas, and the signal strength regression – the rate at which interfering signal power is reduced with increasing distance. Since ground wave and direct wave propagation depend on the nature of the ground covered - urban, sub-urban and rural environments must be studied.

Chapter 4 shows the development of an experimental program that was designed to provide answers to the above points, and in doing so, allow the development of the new generation of EMC standards which can be based on both theoretical models of HF propagation and ‘real world’ measured data.

4 Chapter 4 - Experimental Method

This Chapter describes the practical experiments undertaken by the author to investigate the regression characteristics of electro-magnetic emissions from high frequency 'in building' powerline communication networks. This series of experiments was designed to investigate the near field and far field radiation characteristics of typical 'in building' mains power networks, in urban, suburban and rural environments, when excited by a high frequency signal.

In particular, regression characteristics at 7 MHz, 14 MHz, 21 MHz and 28 MHz, in urban, sub urban and rural environments, were investigated. The experiments were carried out in three separate locations. Firstly, at the University of Lancaster, a quasi-urban area with multiple story buildings. Subsequently the experiments were carried out at two typical houses, one in suburban environment and one in a rural environment.

Measurements were available from experiments carried out within the Lancaster University campus during October 1998. These results were re-interpreted in order to provide an initial indication of regression characteristics, and to aid in the design of the new experiments.

Experiments were carried also out in December 2000 (Lancaster University Campus) and January 2001 (Lancaster University Campus, a suburban house, (Garstang, Lancashire) and a rural house, (Kendal, Cumbria)).

The results from the 1998 experiments were examined intensively in order to provide an initial indication of regression characteristics, and to aid in the design of the 2000/2001 experiments.

4.1 The Lancaster University (1998) Experimental method.

4.1.1 Signal Injection.

A number of buildings were selected as containing suitable signal injection points. The primary building selection criteria was the distance between the building and the wideband HF antenna. It was necessary to obtain field strength figures at a range of distances covering the near field and far field regions surrounding the HF excited mains network. The selection of suitable buildings for the experimental program was also based on accessibility.

At each selected building and signal injection point, the coupling unit was plugged into a standard 13A socket. The power lead of the signal generator was connected via the coupling unit's 'pass through' socket. Once the correct frequency and power level had been selected, the signal generator's output terminal was connected to the coupling unit, via a transient suppresser. Thus, the HF signal was injected onto the power network in unbalanced mode, between live and earth. This arrangement is shown in Figure 30 below.

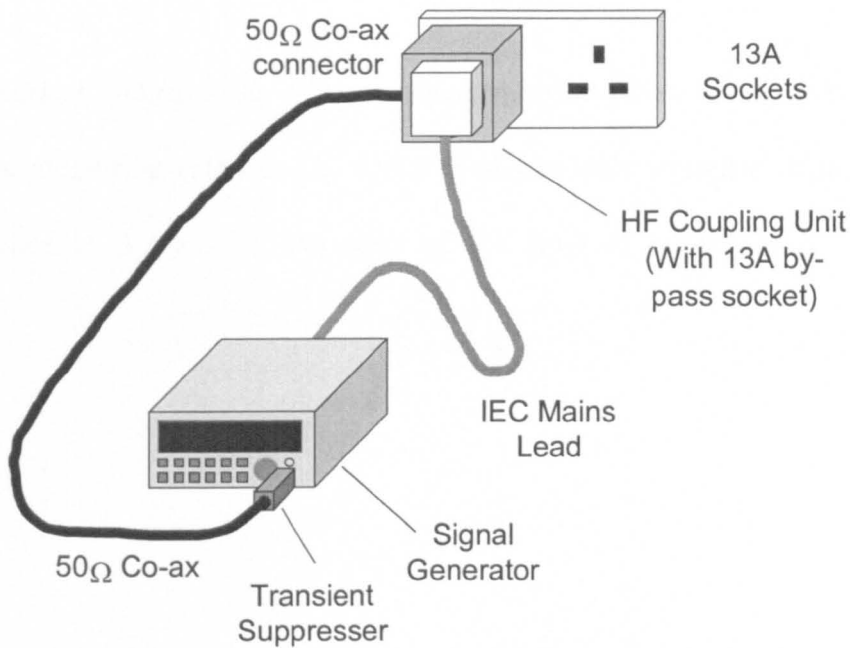


Figure 30 - The Phase 1 HF signal injection arrangement

Co-ordination between the operator of the signal generation / injection equipment and the operator of the receiving equipment was achieved by using mobile telephones.

Initially, a frequency of 3.5 MHz was chosen for the experiment, but after initial measurements, during which the noise floor at 3.5 MHz was found to be approximately -6 dBμV/m, it was decided that there was too much interference at this frequency. The experimental program was continued using 7 MHz.

The signal injection powers selected were +20 dBm, +10 dBm, 0 dBm, -10 dBm and -20 dBm. Where the signal injection point was close to the wideband HF antenna, less than 100 m, only the latter three power levels were injected. The power spectral density (in 9 kHz) for each power level has been calculated. Signals were injected in a live to earth configuration.

Figure 31, overleaf, indicates the selected signal injection points, along with their positions relative to the measuring antenna. The actual distances were measured using a copy of this diagram, printed at A1 width. An copy of the diagram printed at A3 is included in Appendix A.

Ordnance Survey
Superplan Data

Option 1
Site-centred



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This Superplan product does not contain all relevant map information.

20 0 20 40 60 80 100



100 meters

This product contains data surveyed at one or more of the following scales: 1:1250, 1:2500 & 1:10000.

National Grid sheet reference at centre of this Superplan: 554807.

The representation of a road, track or path is an indication of a right of way. Heights are given in metres above datum. The alignment of tunnels is approximate.

Feature Coordinates: 48486, 55254
Supplied by: Western Land Gazette
Serial Number: 74107404

Figure 31 - The Lancaster University Campus

4.2 Signal Measurement.

The electrical field strength, resulting from each injected signal was captured using the wideband HF antenna. This antenna is situated on the roof of the Engineering building, towards the southern end of the university campus.

The feed-line from the antenna was accessible from the laboratory. This is where the ICOM transceiver, antenna tuning unit (ATU) and power supply unit (PSU) were situated. The antenna and feed-line network were tuned to resonance at the defined frequency using the ICOM ATU. This frequency was initially 3.5 MHz, but was subsequently changed to 7 MHz. The ICOM transceiver was configured in the following way :

OPTION	SELECTED
MODE	CW (2.7kHz BW)
NB	OUT
ATT	OUT
PBT	CENTER
NOTCH	OUT
PIT	OFF
AGC	OFF
PRE AMP	ON

Table 4 - ICOM Transceiver configuration

Thus the additional filtering and signal enhancement capabilities of the receiver were switched off and a simple Carrier Wave mode with a 2.7kHz signal bandwidth was used. Noise Blocking, ATTenuation, NOTCH filtering, Pass Band Tuning, PITch modification or Automatic Gain Control.

For each combination of injected signal parameters, (building, injection point and power level), signal meter reading from the ICOM transceiver was recorded. The measurement equipment arrangement is shown in Figure 32 below.

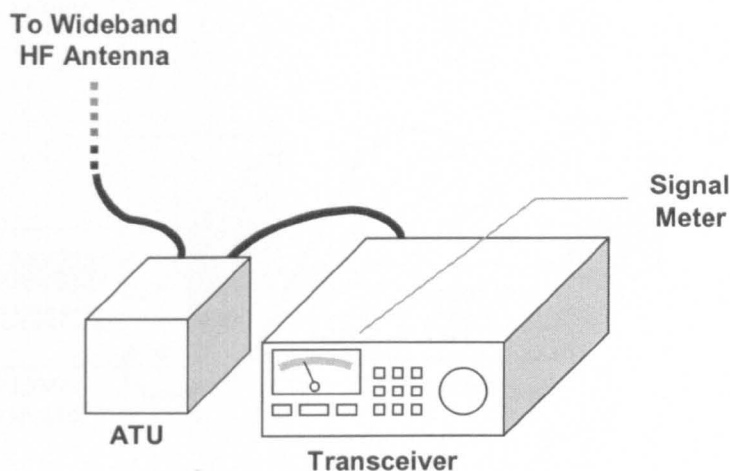


Figure 32 - The 1998 signal measurement equipment arrangement

4.2.1 Receiver 'Calibration'

In order to translate the signal meter readings to true power readings, which could be used to calculate field strength, it was necessary to compile a 'signal meter to dBm' conversion table. This process was similar in nature to a calibration process, although no actual adjustment was made to the transceiver's signal meter.

Before the process was begun, the transceiver was configured exactly as it was during the experimental activity, i.e. as shown in Table 4.

This conversion table was compiled by connecting the receiver directly to a NAMAS calibrated signal generator. With the frequency set to 7 MHz, the power level from the signal generator was adjusted such that for each level on the transceiver's signal meter, an equivalent power level in dBm could be recorded. The equipment arrangement for this process is shown below.

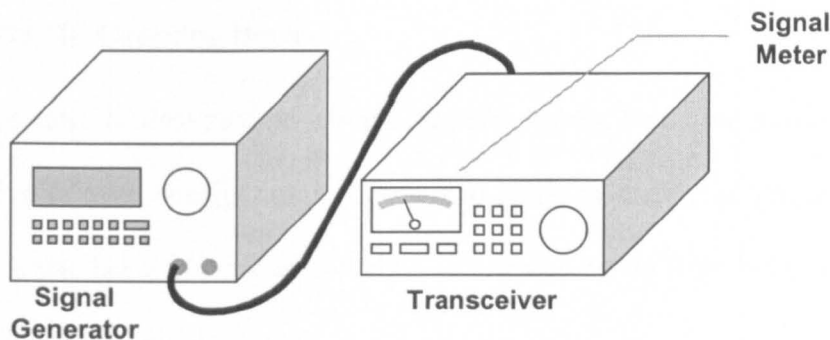


Figure 33 - Equipment arrangement for calibration of the signal meter

The signal meter to dBm conversion table is shown in Table 5 below.

Signal Meter	Power level (dBm)
S0	-111
S1	-109
S2	-106
S3	-103
S4	-100
S5	-97.5
S6	-94
S7	-91
S8	-87
S9	-83

Table 5 - Signal meter to dBm conversion chart for the ICOM IC735 (at 7MHz)

Using the S meter to dBm conversion chart shown in Table 5, it was possible to record signal power levels (in dBm) at the transceiver, along with location and injected signal power information for each signal injection point shown in Figure 31 were recorded. The results of this series of experiments are shown within Chapter 5 of this thesis.

4.3 The 1998 HF Coupling Device

This coupling unit is designed to couple signals safely onto the mains network in an unbalanced live to earth configuration. In order to facilitate testing at single socket injection points, the unit also has a filtered un-switched 13A socket on its front face. This arrangement is shown in Figure 34, below.

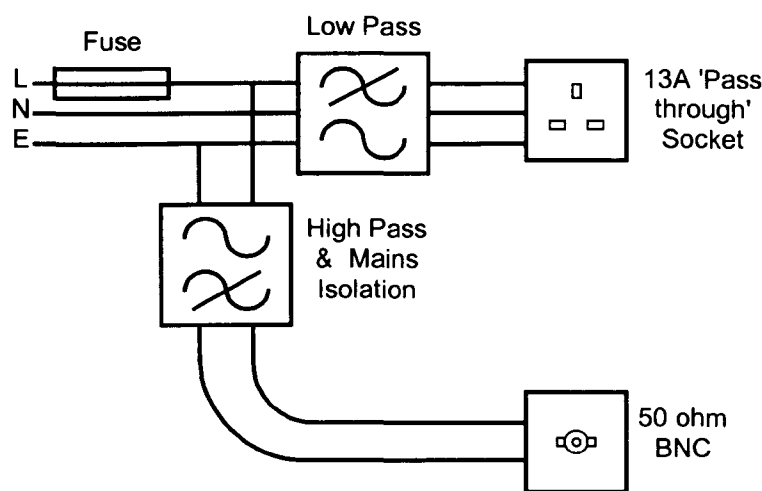


Figure 34 - The Phase 1 HF Coupling unit - internal arrangement

4.3.1 Transient Suppressor

This device, not part of the coupling device but shown in Figure 30, is simply a varistor connected between the inner conductor and the co-ax sheath (earth). The device is contained within a BNC (male) to BNC (female) 'pass through' enclosure.

4.3.2 Wideband HF Antenna

The antenna used throughout this experiment is situated on the roof of the Lancaster University Communication Research Centre (LCRC) at a height of approximately 15m. It is designated FWB/2530 by its supplier, SMC Communications. The specifications are shown in Table 6.

Parameter	Specification
Freq. Range	2.5 - 30 MHz
Power Rating	1 kW Average
Max VSWR	2.5:1
Gain	Up to 3 dBi
Input Impedance	50 Ohms unbalanced
Construction	Heavy duty cadmium copper, with galvanised steel and stainless steel fittings

Table 6 - Wideband HF antenna parameters

The physical geometry and radiation characteristics, taken from the manufacturers supplied data sheet, are shown in Figure 35 and Figure 36, below [41].

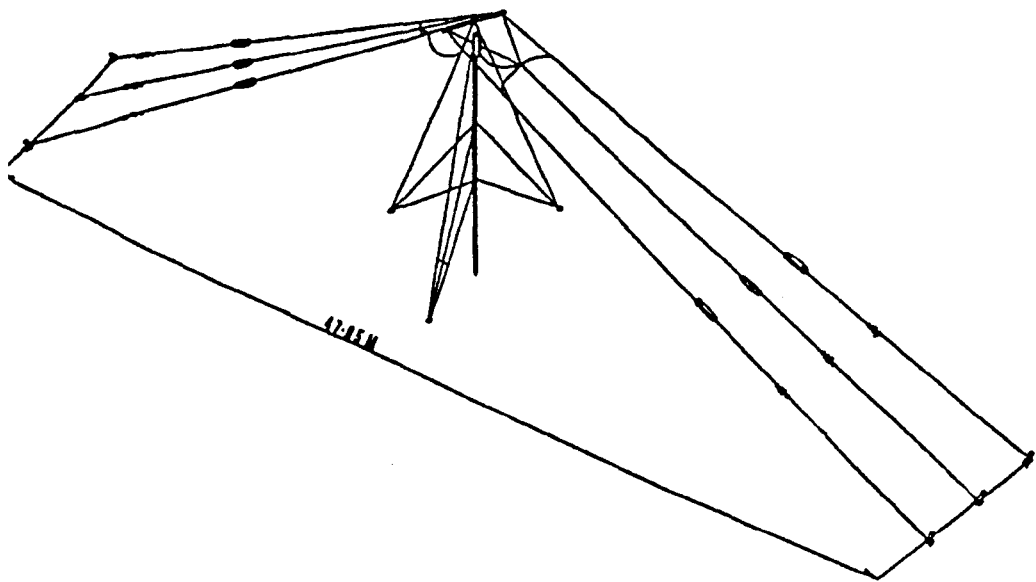


Figure 35 - The HF wideband Antenna

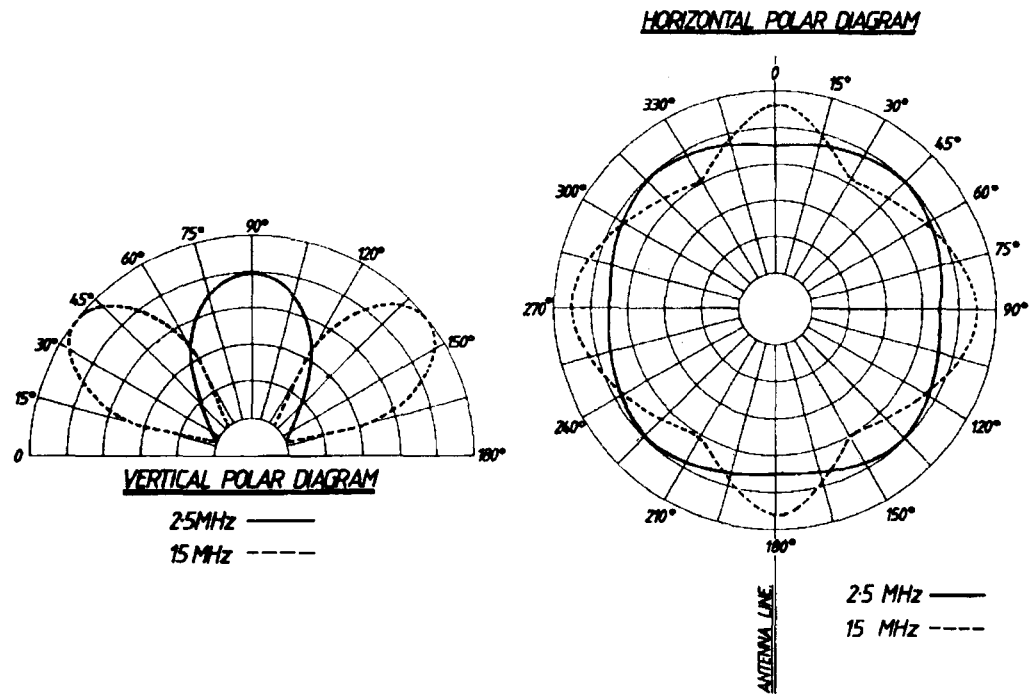


Figure 36 - The wideband HF antenna radiation characteristics

4.3.3 Power Spectral Density of injected signal

For each injected power level, the power spectral density (PSD) of a CW signal in 9 kHz bandwidth has been calculated – see Table 7.

Power (dBm)	PSD (dBm/Hz)
-20	-59.54
-10	-49.54
0	-39.54
+10	-29.54
+20	-19.54

Table 7 - Injected signal power to PSD conversion (in 9 kHz)

Above assumes no loss within the coupling unit, and no impedance mismatch between the coupling unit and mains network at the signal injection point.

4.4 The Urban / Suburban / Rural (2000/2001) Measurement Method.

4.4.1 Introduction.

In these experiments, it was decided to take a different approach to measuring field strength regression. The signal was injected onto the building wiring at one location, or at several locations within one room. A portable antenna was then deployed at a range of distances from the signal injection point, and received signal power was again measured.

4.4.2 Signal Injection.

At each selected signal injection point, the coupling unit was plugged into a standard 13A socket. Once the correct frequency and power level had been selected in the signal generator, the output terminal was connected to the coupling unit. This arrangement is shown Figure 37, below.

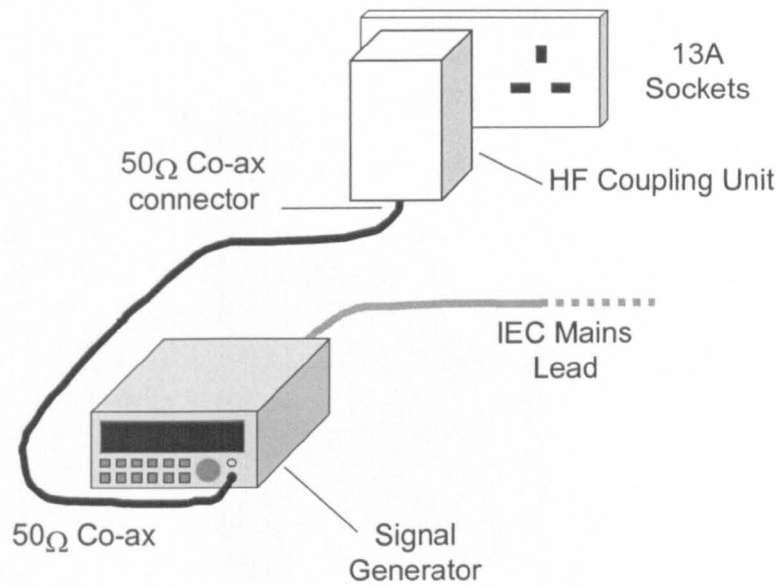


Figure 37 – The Phase 2 HF signal injection arrangement

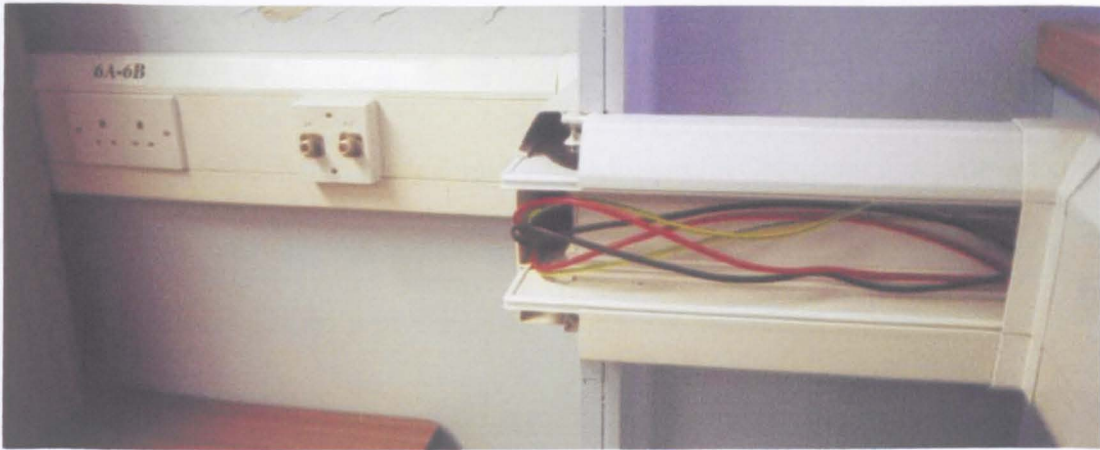
For the initial experiment, the signal was injected in an unbalanced, live to earth mode. This design of coupling unit is shown in photograph 1, overleaf. With all subsequent experiments, the signal was injected between live and neutral. This was considered to be more representative of a likely commercial PLC system. The 2001/02 results presented in appendix 2, indicate socket impedances and phase angles, both for the sockets used in these experiments, and a more general survey.



Photograph 1 - The coupling unit in socket 2B, configured for live - earth signal coupling

Co-ordination between the operator of the signal generation / injection equipment and the operator of the receiving equipment was achieved by using mobile telephones.

In the 2000/2001 part of the experimental program, the regression at 7 MHz was re-investigated. This time, the signal was injected at a power level of 0 dBm, using the new coupling unit with live to earth signal coupling. Sockets 2A, 3B, 4B, 5A, 6B, 9A, 8A#, 7A# within the power line communications lab were used as signal injection points. As a comparison, the 0 dBm signal was also injected onto the wideband antenna, which is situated directly above the power line communications lab. The wiring arrangement in the powerline lab is shown in photograph 2, below.



Photograph 2 - The mains wiring arrangement within the powerline lab at Lancaster University

On a subsequent date, regression characteristics at 14 MHz, 21 MHz and 28 MHz were investigated. This time, the injected signal power was +10 dBm, using a coupling unit configured for live to neutral signal coupling. The increase was necessary in order to obtain measurements over a greater distance, and at higher frequency than before.

An experiment was undertaken at a typical suburban house in Garstang, Lancashire. In this experiment, the regression characteristics at 14 MHz, 21 MHz and 28 MHz were investigated, using +10 dBm injected signal power.

Similar experiments were also undertaken at a rural house in Kendal, Cumbria, into the regression characteristics at 21 MHz with +10 dBm injected signal power. The regression profile from mains signal injection was compared with that of a tuned indoor dipole at the same signal power and location.

4.4.3 Signal Measurement.

For signal reception at 7 MHz, an accurately tuned, portable, inverted 'v' dipole antenna was constructed. This antenna was approximately 6.5 meters tall at its highest point. This antenna was deployed in two positions, 140 m and 220 m distant from the signal injection point within the laboratory.

On subsequent occasions, a multi-band G-whip antenna, attached to the roof of a four wheel drive vehicle was used. Signal strength and position measurements, (from a GPS navigation system), were made in a range of locations surrounding the signal injection point. The receiving arrangement is shown in Figure 38, and in Photographs 3 – 7, below.

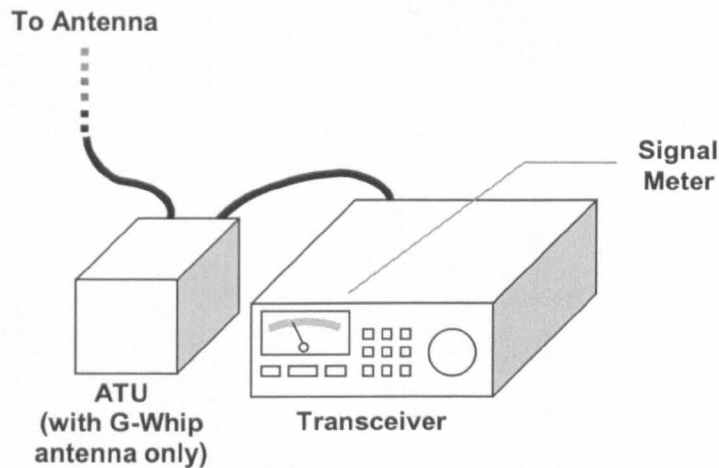


Figure 38 - The signal measurement equipment arrangement



Photograph 3 - The HF-225 receiver along with battery pack and laptop computer



Photograph 4 - HF-225 receiver and battery pack at the base of the inverted V antenna mast



Photograph 5 - The inverted V dipole antenna attached to the four wheel drive vehicle



Photograph 6 - The Inverted V dipole antenna at position 1, on 27th December 2000 at
Lancaster University.



Photograph 7 - The inverted G whip antenna attached to the four wheel drive vehicle

4.4.4 Receiver 'Calibration'

As before, it was necessary to compile a 'signal meter to dBm' conversion table. This conversion table, shown in Table 8 below, was compiled by connecting the receiver directly to a signal generator as shown below. With the frequency set, the power level from the signal generator was adjusted such that for each level on the transceiver's signal meter, an equivalent power level in dBm could be recorded. The 2000/2001 receiver calibration setup is shown in photograph 8, overleaf.



Photograph 8 - The receiver 'calibration' equipment set-up

Signal Meter	Detected Power (dBm)			
	7 MHz	14 MHz	21 MHz	28 MHz
S1	Noise	Noise	Noise	Noise
S2	-105.4	-105.2	-106.3	-105.3
S3	-101.9	-102.1	-103.1	-101.5
S4	-98.7	-98.6	-100.2	-98.8
S5	-95.3	-95.5	-97.0	-95.3
S6	-91.3	-90.9	-92.3	-91.0
S7	-86.1	-86.0	-88.1	-86.1
S8	-80.7	-81.0	-82.8	-80.6
S9	-74.8	-74.9	-77.0	-74.6
S9+10	-60.8	-61.3	-62.9	-61.2
S9+30	-49.1	-49.6	-50.7	-49.2
S9+50	-27.2	-27.9	-27.9	-27.3

Table 8 - Signal meter to dBm conversion chart for HF-225 Receiver

With the results of both groups of experiments, it was necessary to calculate the field strength at the antenna, given a known signal power at the transceiver. The following relationship, derived from Equation 6, but including an estimated antenna and feedline loss factor of 0.5 as described at the end of chapter 3, was examined.

$$P_R = \frac{G_R}{960} \left(\frac{\lambda E}{\pi} \right)^2 \dots\dots\dots \text{Equation 7}$$

In Equation 7 above, P_R is the received signal power. G_R is the numerical gain of the receiving antenna (in the direction of the signal source), over a theoretical isotropic antenna. E is the signal field strength at the antenna.

In this case, the gain, G_R , of the wideband HF antenna was assumed to be that of a standard dipole, i.e. 2.15 dBi, or in numerical terms, 1.64.

By re-arranging Equation 7 :

$$E = \frac{\pi}{\lambda} \sqrt{\frac{960 P_R}{G_R}} \dots\dots\dots \text{Equation 8}$$

Equation 8 was used to calculate the field strength in $\mu\text{V/m}$ and subsequently in $\text{dB}\mu\text{V/m}$, i.e. expressed as a dB ratio relative to a field strength of 1 $\mu\text{V/m}$.

For each experiment, the field strength in $\text{dB}\mu\text{V/m}$ was plotted against distance between the signal injection point and the antenna.

4.4.5 Antenna / Feedline Voltage Standing Wave Ratios (VSWRs)

For each receiving antenna / feedline combination, the VSWR was measured and recorded in polar format, i.e. modulus of impedance in Ohms and phase angle in degrees.

Description	Frequency (MHz)	VSWR	Impedance (Ohms)	Angle (Deg.)
Phase 2 – Lancaster Position 1	7	1.42:1	35	-6
Phase 2 - Lancaster Position 2	7	1.34:1	37	-9
Phase 2 - Lancaster	14	2.20:1	25	20
Phase 2 - Lancaster	21	1.47:1	34	-11
Phase 2 - Lancaster	28	2.64:1	123	19
Phase 2 - Garstang	14	2.78:1	27	42
Phase 2 - Garstang	21	3.55:1	20	40
Phase 2 - Garstang	28	3.00:1	145	6
Phase 2 - Kendal	21	1.65:1	31	15

Table 9 - The VSWR for each receiving antenna / feedline combination

4.5 Measured Socket Impedances

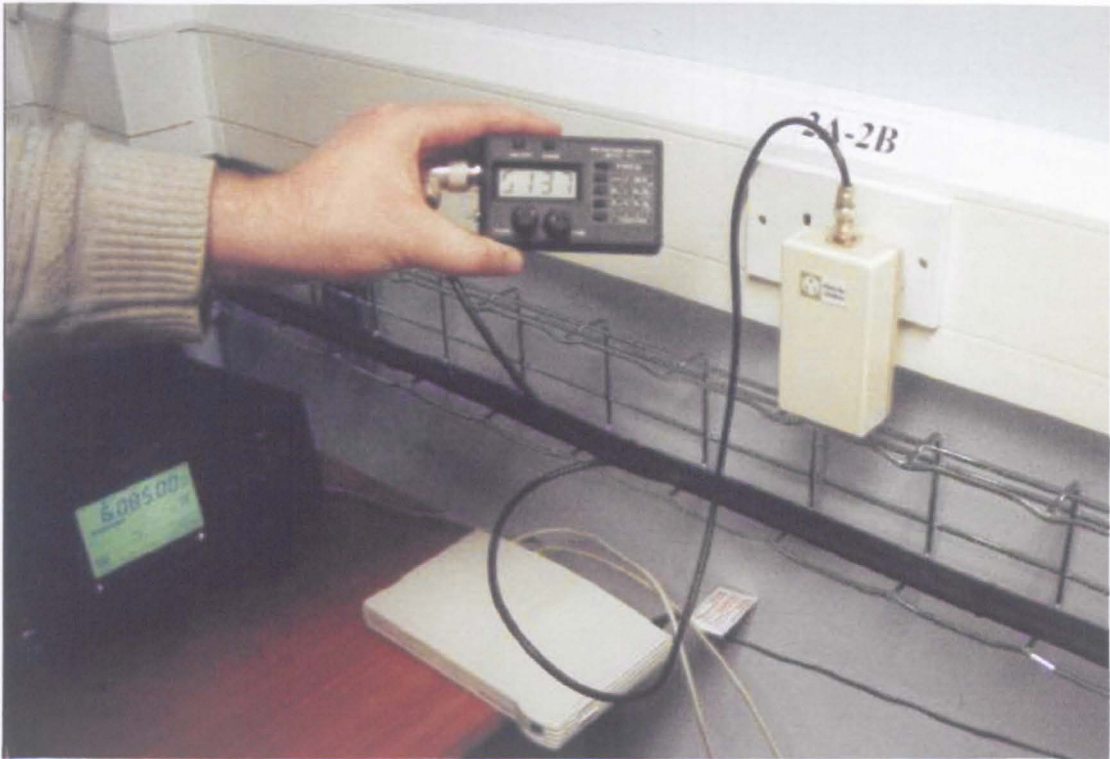
As part of the 2000/2001 experimental program, it was necessary to measure the impedance at every frequency / socket / coupling unit combination that was used for signal injection. This was achieved using a measuring device originally developed for antenna / feedline evaluation.

In each case the device was tuned to the frequency of interest, after which impedance (modulus) and angle could simply be read off in ohms and degrees and recorded. Since measurement was straightforward, a more general survey was undertaken. All results are presented in the next chapter of this thesis.

A wide range of impedances were observed, mostly between 30 Ω and 100 Ω , but with some deviation beyond this. Photographs 9 and 10 overleaf illustrate the impedance measurement equipment.



Photograph 9 - The Impedance, Angle and VSWR measuring device (with 50 Ω terminator)



Photograph 10 - The impedance measurement device and WBS coupling unit in action

4.5.1 The effect of impedance mismatch

It is clear from the information above that there will be a significant impedance mis-match between the 50 Ω signal source and the mains network coupling unit, and, to a lesser extent, between the receiving antenna and the 50 Ω co-axial feedline and receiver. In both cases a measurable standing wave may be set up on the co-axial cable. It can be seen from Figure 39, below that where the VSWR is 1.5 or less, there is no significant reflection of signal power. Within the context of the investigation of far-field HF regression, with field strengths calculated in dBuV/m, VSWRs of up to 2.5 may be tolerable [15].

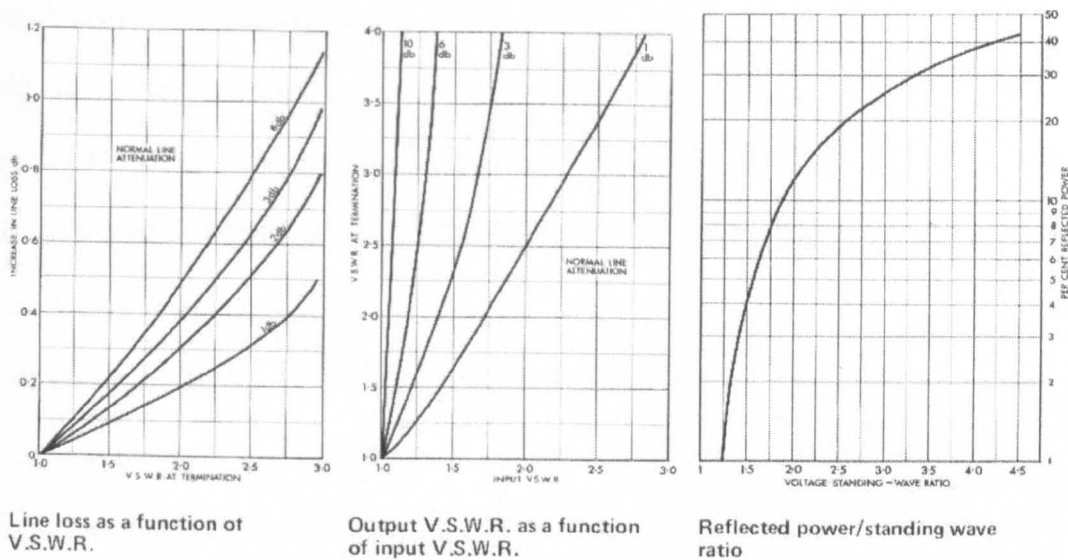


Figure 39 - The effect of VSWR on reflected / absorbed signal power

4.6 Smith Charts

Although it is possible to quantify the effects of impedance mismatch using an algebraic approach, a more intuitive understanding can often be found in one of the many graphical methods. The most commonly used method, was developed by Philip H Smith in 1944. [15] Figure 40 shows a smith chart with resistance axis (central horizontal) and reactance axis (multiple arcs centred on a vertical line through position 0,0 i.e. at the extreme right of the chart). Point a represents an normalised impedance of $2+j0.5$, point b represents $1-j1.2$. [40]

Take point b and drawing a circle centred on position $1+j0$ with a radius matching the distance between $1+j0$ and point b. By extending a line from point b, through the centre of the chart ($1+j0$) to the opposite side of the circle, a new point can be plotted. This point, in this case $0.42+j0.48$, represents the admittance corresponding to the impedance of point b. This is illustrated in Figure 41, overleaf where a load impedance of $0.8+j1$ is plotted. When connected to a transmission line of nominal impedance $1+j0$, the angle of reflection co-efficient can be read by extending a line from the centre of the chart through the load impedance to the outer axis. In this case the angle of reflection co-efficient (ψ) would be 72 degrees. The magnitude of the reflection co-efficient can be read by measuring the horizontal distance between $1+j0$ and the load impedance 'd'. The figure can be read directly from the uppermost of the 4 scales at the bottom of the chart. In this case the reflection co-efficient magnitude is 0.15

In the last example, Figure 42, the VSWR can be derived by plotting a circle as before centred on $1+j0$ and reading from the lower left hand scale (in fact the resistance scale yields the same figure) In this case the VSWR is 3.0.

Chapter 4

NAME	TITLE	DWG. NO.
SMITH CHART FORM 5301-7560-N	GENERAL RADIO COMPANY, WEST CONCORD, MASSACHUSETTS	DATE

IMPEDANCE OR ADMITTANCE COORDINATES

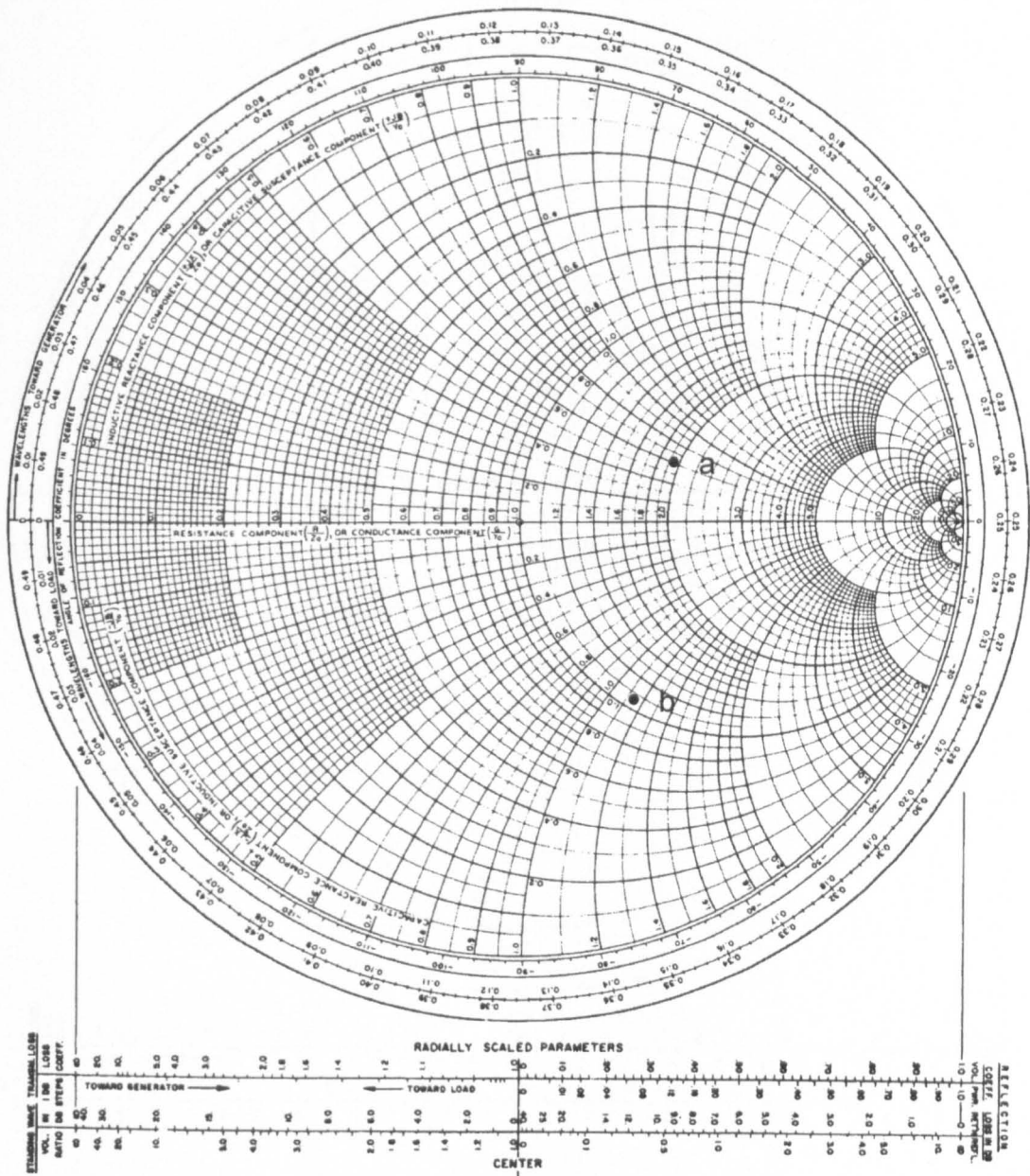


Figure 40 - Smith Chart with normalised impedances a and b plotted [40]

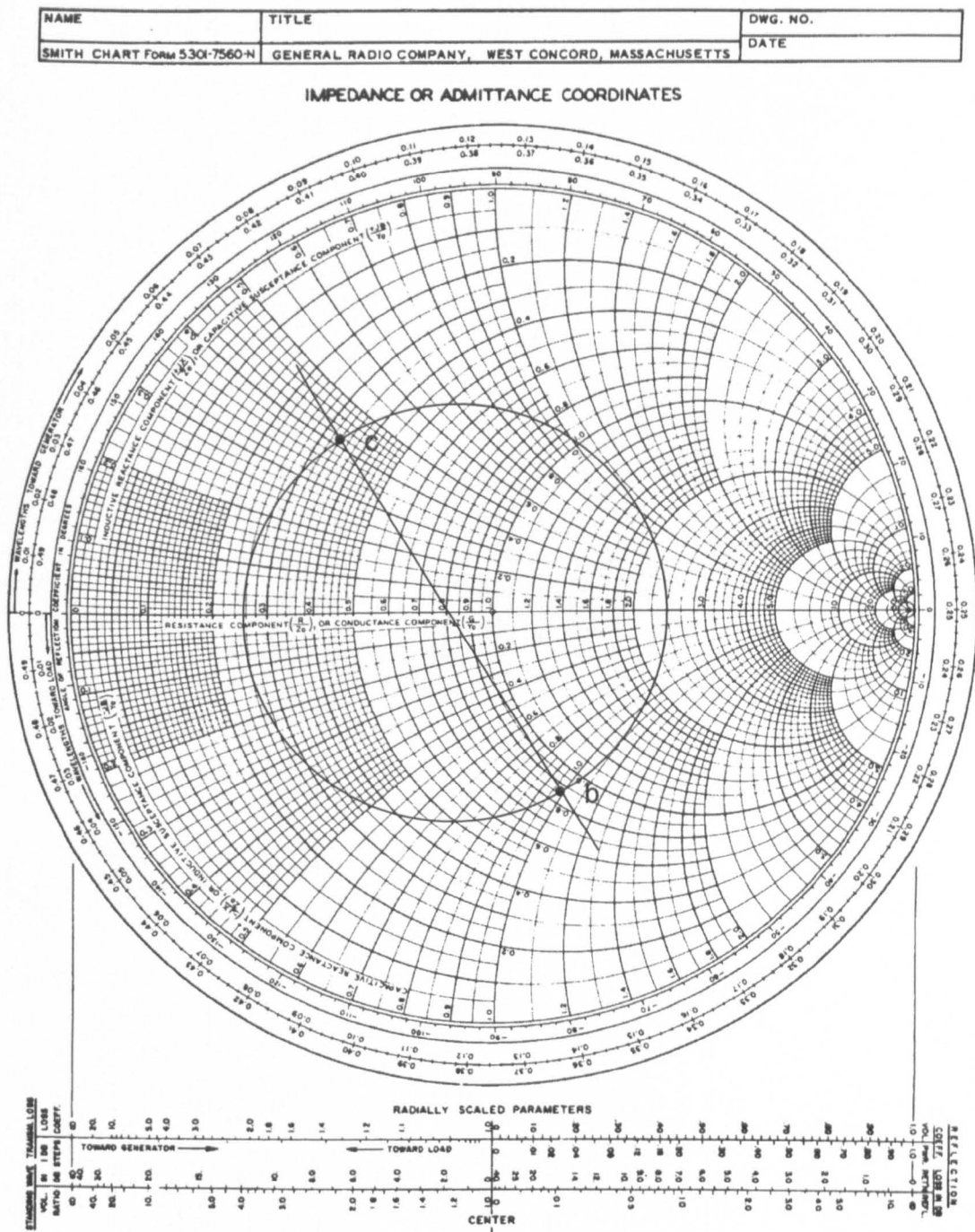


Figure 41 – Impedance at point ‘b’ shown and admittance ‘c’ derived [40]

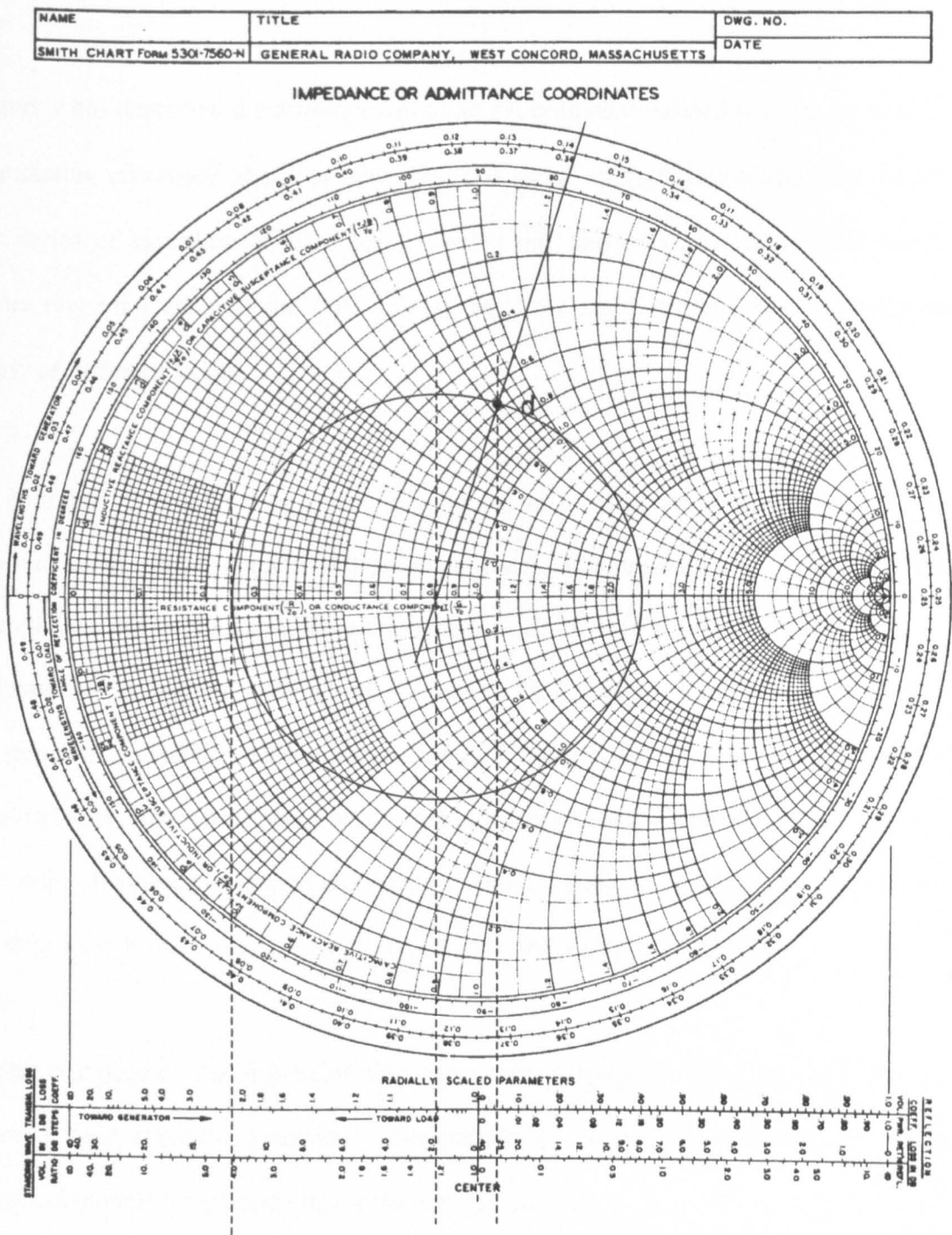


Figure 42 - Load impedance 'd' on a transmission line [40]

4.7 Chapter 4 conclusions

Chapter 4 has described the development of an experimental method that can be used to assess the radiating efficiency of power networks, and signal strength regression. By the time the later series of experiments were being carried out, the experimental method deployed a mobile receiving antenna and GPS system to allow rapid surveys to be undertaken in a variety of different locations.

The later part of chapter 4 also examines the effect of impedance mismatch at the point of signal injection. In the context of these experiments, this problem has the effect of reducing the radiating efficiency, and therefore antenna factor of the mains network, and can be accounted for in that way. With high bandwidth PLT systems, it is necessary to achieve the best impedance match possible between the PLT modem and coupling arrangement and the mains network, to achieve maximum signal power transfer and therefore maximum signal to noise ratio. However, as shown in the next chapter, significant differences in signal coupling impedance can be seen when multiple signal injection points are examined.

Chapter 5 discusses the results of this experimental program, focusing especially on the observed field strength regression. Comparison is made with a number of established theoretical models for ground / direct wave propagation.

5 Chapter 5 – Results

The previous chapter developed an experimental method for the evaluation of high frequency radiated emissions from PLC networks. This chapter describes the results obtained from both the 1998 (phase 1) experiments, and the more comprehensive 2000/2001 (phase 2) experiments.

During this time, the experimental technique was evolved from a fixed antenna / mobile signal injection point arrangement, to a fixed injection point / mobile antenna arrangement. This second arrangement used during the 2000/2001 experiments was able to yield more accurate results by virtue of its repeatability, and the number of results that could be obtained.

It was decided not to use injected signal powers higher than +20 dBm, since there was a danger of interference to radio services. The 7.0 MHz and 3.5 MHz frequency bands are used by long distance amateur radio. In these frequency bands, transmitting +20 dB (0.1 W) via a mostly ineffective antenna (i.e. the mains network) within a private site (i.e. the university campus) should not cause interference. If the injected signal power was increased beyond +20 dBm, even considering the non ideal antenna characteristics of the mains network, there is a danger of causing interference to amateur radio users who may be attempting to capture weak signals from distant transmitters.

5.1 Results from the 1998 experiments

The results gathered from the 1998 experiments are shown in Table 11 to Table 16, in Appendix 2. The subsequently presented charts, Figure 46 to Figure 50 indicate measured electrical field strength over a number of discrete distances. With these experiments, it is the signal injection point that is being moved. Thus there is also some variation in injected signal power associated with each signal injection point. This can be clearly seen in the two field strengths measured from signal injection at the Old and New Library buildings (Figures 46 to 50). In all cases, there is a 6 dB μ V/m difference in the observed field strength, despite the fact that the distances between injection point and receiving antenna are identical. It is reasonable to assume a similar degree of variation could be present with all signal injection points.

Another drawback with the 1998 results becomes evident when the results of signal injection at -10 dBm and -20 dBm are examined. At these signal injection powers, it is not realistically possible to derive any firm conclusions as regards regression, since, although the received signals can be heard, the actual received signal power and therefore the calculated field strength is within the radio noise floor.

The electromagnetic noise floor at the site of the university antenna was approximately -15 dB μ V/m at 3.5 MHz and 7 MHz in the receiver's 9kHz bandwidth. From the corresponding charts (for signal injection at -10 and -20 dBm) it is clearly difficult to show regression beyond approx 100 meters.

From the results obtained at +20 dBm and +10 dBm (Figures 46 and 47) a field strength regression of 20 dB per decade of distance can be demonstrated by manually fitting a straight

line to the observed field strengths in dB μ V/m plotted on a logarithmic distance scale. The accuracy of this figure is however de-graded by the limited number of injection points, and by the observed field strengths beyond 200m being close to the radio noise floor.

5.2 Further Notes from 1998 experiment.

5.2.1 Interfering carrier

The received signal power resulting from signal injection at Furness College Lecture Theatre No.1 was affected by an interfering signal. This signal, noted in Table 11, was noticed when the injected signal was switched off, caused a signal meter reading on the receiver of S2, or -106 dBm. Both signals were measured within the receiver's input bandwidth (in this configuration) of 2.7 kHz.

In order to compensate for this interference, the signal meter reading from the transceiver, used to calculate field strength, was reduced by S2 for each reading from this location and time.

5.2.2 Trend lines

Trend lines have been added to each of the charts (Figures 46 to 50) that plot E field strength against a logarithmic distance scale. In the manual positioning of these lines, greater consideration has been given to fitting the near field (less than 180m) points. At greater distances the measured signal power, was at or below the lower measurement limit of S0, or

–111 dBm on the transceiver. This has the effect of artificially raising the calculated field strength at these points.

These lines reflect the correct shape for electric field strength regression, given below.

$$E = w \left(\frac{x}{d^3} + \frac{y}{d^2} + \frac{z}{d} \right)$$

Equation 9 - A general expression for field strength regression

In the above equation, d is the distance from the conductor, and w , x , y and z are constants.

5.2.3 *K factor*

It is often useful when assessing complex antenna designs, to evaluate the receiving efficiency of a particular antenna design. The antenna factor or K factor can be calculated as the ratio between incident electrical field strength (V/m) to voltage at the antenna terminals (V).

The K factor is therefore expressed in units of m^{-1} , or in dB terms dB/m.

Conventionally enhanced antenna designs are highly directional, so have large K factors. Power line networks, due to their complex, but in RF terms, random configuration can be expected to be roughly isotropic (non directional) in nature with a much smaller K factor.

For each combination of injected signal power and injection point, the K factor of the power network was calculated by comparison to the wide-band antenna directly above it, on the roof of the laboratory. The peak value, found with an injection point 18m from the antenna position, was -84 dB/m, (at 7MHz and 0dBm injected signal power). At this location, (Room B35 engineering in Table 14, Table 15 and Table 16, the correlation between K factors measured at 0 dBm, -10 dBm and -20 dBm injected signal power was particularly good at a deviation of only 1dB/m.

Taking those signal injection points from Table 14 that were less than 180m from the receiving antenna (within the near field), the calculated K factor in each case showed a deviation of no more than 7 dB/m across the injected power range 0 dBm to -20 dBm.

At distances greater than 180 m there is significant deviation in K factor results for given injection points. Again, this has occurred because the measured signal power was at or below the lower measurement limit of S0, or -111 dBm on the transceiver. Thus the measured change in field strength for a given change in injected signal power could not be accurately reflected in the K factor calculation.

5.3 Results from the 2000/2001 experiments

The results from the 2000/2001 experiments are shown in Table 17 to Table 25 in Appendix 3.

Table 17, along with the subsequent chart, figure 51, show the 7 MHz experiments carried out using the inverted V dipole antenna, tuned to exactly 7MHz, in two positions within the University Campus. In each receiving antenna position, a single frequency 0 dBm signal was injected into each socket within the power line laboratory in turn, and the received signal power was recorded. In some cases, there was no signal detection with an injection power of 0 dBm. In these cases, highlighted in Table 17, the signal power was increased to the figure shown and the resulting field strength extrapolated to correspond with a 0 dBm injected signal power.

In this case there were only two receiver positions, and the intention was not to derive regression information, but to compare the effects of injecting into different points on the same LV network. As expected, the chart in figure 51 shows a correlation between field strengths from different signal injection points on the same LV network. In both cases, the deviation is no more than 10dB μ V/m. Measured field strengths were, on average, higher at the second antenna position, 220m from the power line lab. The exact reason for this might have become clear if more antenna positions had been investigated but it was decided to move to a more portable receiver setup at higher frequency.

Table 18, Table 19 and Table 20 in appendix 3, show measured field strengths at 14 MHz, 21 MHz and 28 MHz respectively from a large number of receiving points at known grid

references, and hence known distances from a single injection point in the power line lab. In these cases, the large number of points allow a clear regression trend to be drawn. For the Lancaster University campus, which might be described as an urban environment, the regression figures indicated for 14, 21 and 28 MHz are -40 dB / decade, -41 dB / decade and -42 dB / decade of distance respectively.

Similar experiments in a sub-urban house in Garstang, Lancashire, from Table 21, Table 22 and Table 23, yielded a similar figures of -38 , -39 and -40 dB/decade for 14, 21 and 28 MHz respectively.

In a rural environment in Kendal, Cumbria, from Table 24 and Table 25 the figure at 21 MHz was -35 dB/decade, the lowest regression figure observed. In a rural environment, although ground wave propagation is likely to be similar to urban and sub-urban environments, there are fewer physical obstacles between signal injection point and receiving antenna. Hence the lower regression figure.

5.4 Further Notes from 2000/2001 experiments.

5.4.1 Trendlines

The trend lines, shown on each 2000/2001 chart as a broken line, have been manually applied to each field strength chart - figures 51 to 59. In this case, the each data point was given equal consideration when the trend line was manually positioned.

5.4.2 Comparison with existing VHF signal propagation models

In evaluating the observed field strength regression, a number of existing models have been examined. The author is not aware of any existing models covering field strength regression in the HF band, beyond the simple free space model used to predict direct wave and reflected wave (ionospheric) propagation over large distances.

Several models do exist however covering short range (in the order of 10 km) VHF propagation. These models have been developed over the past few decades in order to assess the equipment and antenna requirements for operation of two way radio cells in the 30 MHz to 300 MHz frequency band.

In addition to the free space model, expressed simply as a regression of -20dB/decade of distance, the EGLI, Flat Earth, Rough Ground and CCIR VHF propagation models [42] have also been evaluated assuming a transmitter height (the mains network) of 1m and a receiver height (the portable antenna) of 3m. Although it is difficult in practice to define the actual

height of a power network, in fact in some cases its height might be negative, the figures above are felt to be sufficiently accurate for this comparison.

In the case of the rough ground model, a land usage factor is required. Although the land usage within the campus (close to the antenna) was medium / high, in the open fields beyond its outer perimeter, the land usage is close to zero.

The predicted field strength regression for each model is shown in figure 60, alongside measured field strength regression at Lancaster university from the wideband antenna, and the powerline network.

It can be seen in figure 60 that all but one of the models examined overestimated the predicted field strength at 100m by a considerable degree - between 32dB in the case of the Flat Earth model to 50dB in the case of the CCIR model.

The Rough Ground model can be made to fit quite closely to the observed field strength when a land usage factor of 10% is used. Given the geography of the Lancaster University campus, this land usage factor may not be too inaccurate. Ultimately though, the field strength regression per decade appears to be too high when distances greater than 500m are considered.

5.5 Measured Socket Impedances

The results of the socket impedance survey are shown in Table 10, overleaf. The observed impedances were mostly between 30 and 100 Ω , but with some deviation beyond this. These results were measured at a large number of sockets in domestic environments in three

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separate properties. As discussed in the previous chapter, these results quantify the possible impedance mismatches between a powerline modem / mains coupling unit combination, with a characteristic impedance of 50Ω , and various different points on the mains network.

Building	Location	@ 7 MHz	
		[z] (Ohms)	Angle (Deg)
House 1	Upstairs Hall	68	39
Built 1970s	Front Bedroom (W)	64	28
Kendal, Cumbria	Front Bedroom (Wall)	71	0
	Downstairs Hall	43	41
	Kitchen Sk 1	51	28
	Kitchen Sk 2	56	11
	Kitchen Sk 3	27	20
	Kitchen Sk 4	165	21
	Kitchen Sk 5	113	13
	Kitchen Sk 6	51	26
House 2	Cooker	71	33
Built 1990s	Battery Charger	69	10
Kendal, Cumbria	Kettle 1	39	12
	Kettle 2	50	3
	M. Machine	38	36
	B. Cooker	45	-47
	I. Socket	25	7
	D. Table	48	34
	Home PC	83	34
	Study PC	40	37
Powerline Lab	Socket 2B	215	38
Metal Framed Building	Socket 3A	119	10
	Socket 3B	70	26
Lancaster Uni	Socket 4B	55	10
	Socket 5A	45	35
	Socket 5B	36	22
	Socket 6A	102	-21
	Socket 6B	102	-24
	Socket 9A	68	60
	Socket 9B	68	60
	Socket 8A#	65	12
	Socket 8B#	158	-7
	Socket 7A#	83	19
	Socket 7B#	82	19
	Socket 6A#	87	-34

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Building	Location	@ 7 MHz	
		[z] (Ohms)	Angle (Deg)
House 3	Garage, near main door	132	55
Built 1970s	Garage, above bench (bottom left)	40	-43
Garstang, Lancs	Garage, above bench (bottom right)	40	-40
	Garage, above bench (top left)	39	-44
	Garage, above bench (top right)	45	-40
	Hall (left of radiator)	60	-52
	Lounge (behind settee 1)	92	42
	Lounge (behind settee 2)	114	11
	Lounge (under TV 1)	73	19
	Lounge (under left speaker 1)	61	22
	Lounge (under left speaker 2)	57	22
	Lounge (under right speaker 1)	100	9
	Lounge (under right speaker 2)	106	39
	Dining Room (behind cabinet 1)	44	27
	Dining Room (behind bureau 1)	157	18
	Dining Room (behind bureau 2)	122	55
	Kitchen (Computer 1)	101	-34
	Kitchen (Computer 2)	106	-44
	Kitchen (Computer 3)	77	-32
	Kitchen (Computer 4)	76	-28
	Kitchen (Computer 5)	80	-17
	Kitchen (Computer 6)	76	9
	Kitchen (under stereo 1)	126	-6
	Kitchen (under stereo 2)	120	-8
	Kitchen (Coffee machine 1)	50	-46
	Kitchen (Coffee machine 2)	45	-44
	Kitchen (Toaster 1)	70	-42
	Kitchen (Toaster 2)	52	-46
	Kitchen (Microwave 1)	50	-67
	Kitchen (Microwave 2)	39	7
	Kitchen (Mixer 1)	55	25
	Kitchen (Mixer 2)	54	29
	Kitchen (Rice cooker 1)	32	-41
	Kitchen (Rice cooker 2)	32	-39
	Kitchen (Kettle 1)	95	43
	Kitchen (Kettle 2)	85	53
	Kitchen (Next to fridge 1)	156	51
	Kitchen (Next to fridge 2)	163	49
	Landing (Wall Socket)	52	6
	Front Bedroom (next to wardrobe)	136	62
	Front Bedroom (next to bed 2)	110	-55
	Bedroom 1 (under desk 1)	47	-35
	Bedroom 1 (under desk 2)	76	28
	Bedroom 2 (under desk 1)	68	11
	Bedroom 2 (under desk 2)	80	21
	Conservatory 1	49	-39
	Conservatory 2	40	-32

Building	Location	@ 7 MHz	
		[z] (Ohms)	Angle (Deg)
House 4	Kitchen	74	4
Built 1960s	Cooker	34	25
Preston, Lancs	Hall	130	21
	FBedR	56	-32
	BBedR	88	12
	LouCor	103	-5
	KbyUw 1	91	55
	KbyUw 2	139	38
	Garage	85	-39
	KWaM	57	40

Table 10 - Socket Impedances and phase angles.

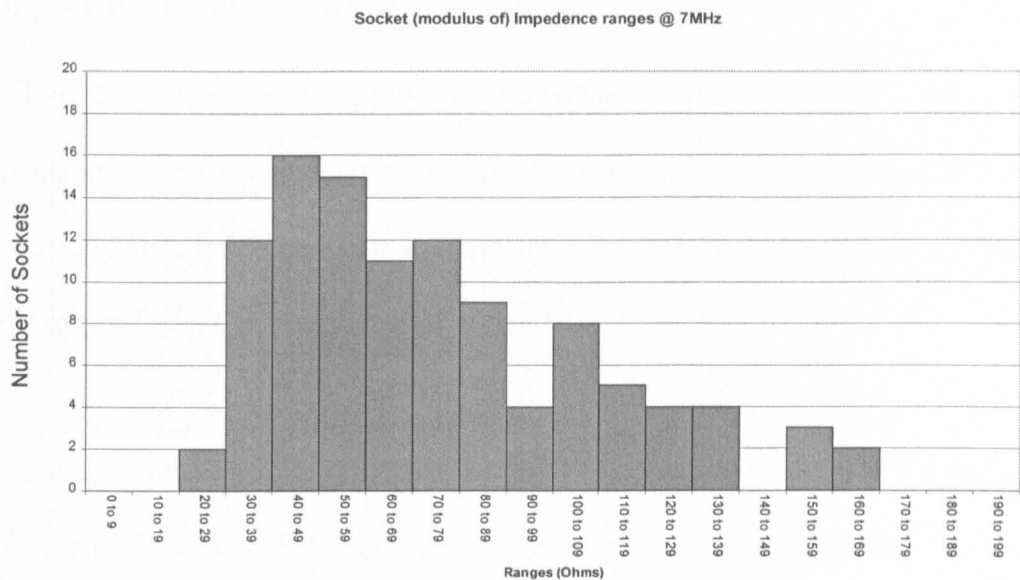


Figure 43 - The impedance (modulus) statistical distribution

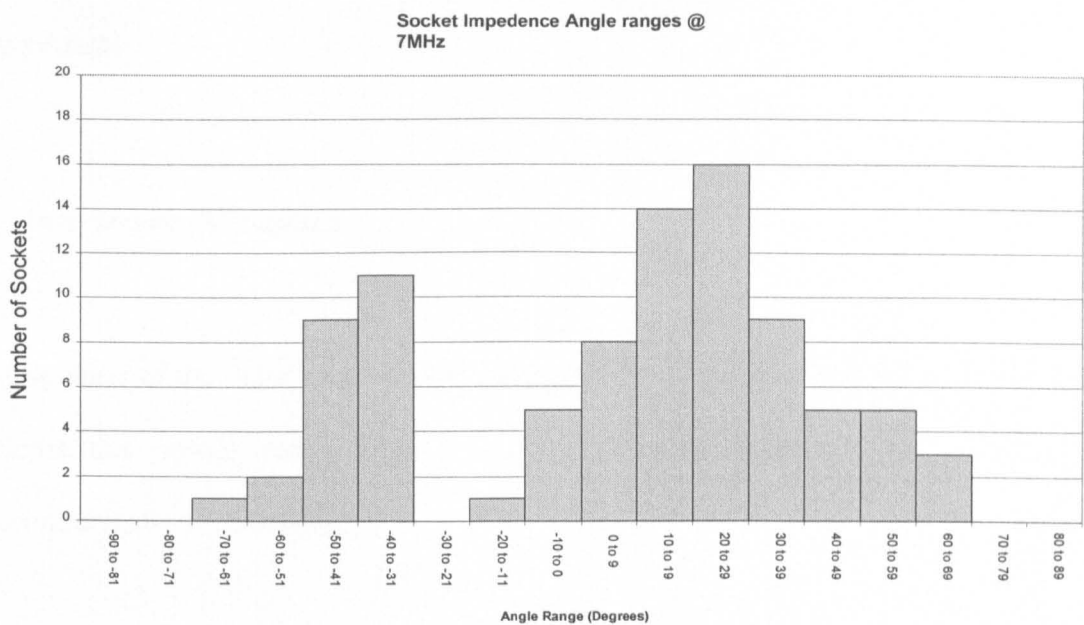


Figure 44 - The impedance (angle) statistical distribution

6 Chapter 6 – Conclusions & Discussions

Chapter 1 of this thesis discussed some of the technical obstacles facing a full scale PLT deployment, and went on to identify electromagnetic interference as one such obstacle. Chapters 2 and 3 detailed some of the existing and proposed standards, CISPR 22, NB30 and MPT1570 respectively, that could be applied to PLC.

The experimental program described in Chapter 4, and results discussed in Chapter 5 and shown in Appendices 2 and 3, have been able to quantify the launched electromagnetic energy at a number of frequencies, for a known power spectral density. It has also been possible to provide a model for field strength regression in Urban, Suburban and Rural environments.

6.1 Interference Mitigation

Existing users of the radio spectrum have legitimately focussed on the potential interference problems that would result from a wide scale commercial roll out of power line telecommunications technology.

As discussed previously there are two scenarios in which electromagnetic interference is possible.

6.1.1 Localised interference

In this scenario, the radio receiving antenna is located in close proximity (a few meters) to a power line network. It can be assumed that conducted emissions into the radio receiver via the power supply can be reduced by efficient filtering at the mains port. Electromagnetic interference is dominated by near field electrostatic (capacitive) and magnetic coupling rather than ‘true’ electro-magnetic radiation. Realistically, this problem can only be avoided by careful frequency selection.

6.1.2 Cumulative ‘at a distance’ interference

Much of the power line EMC debate has revolved around the potential cumulative effects of large numbers of PLC systems. Although non-coherent, individual PLC systems may contribute to a significant rise in the noise floor over a large distance. [43]

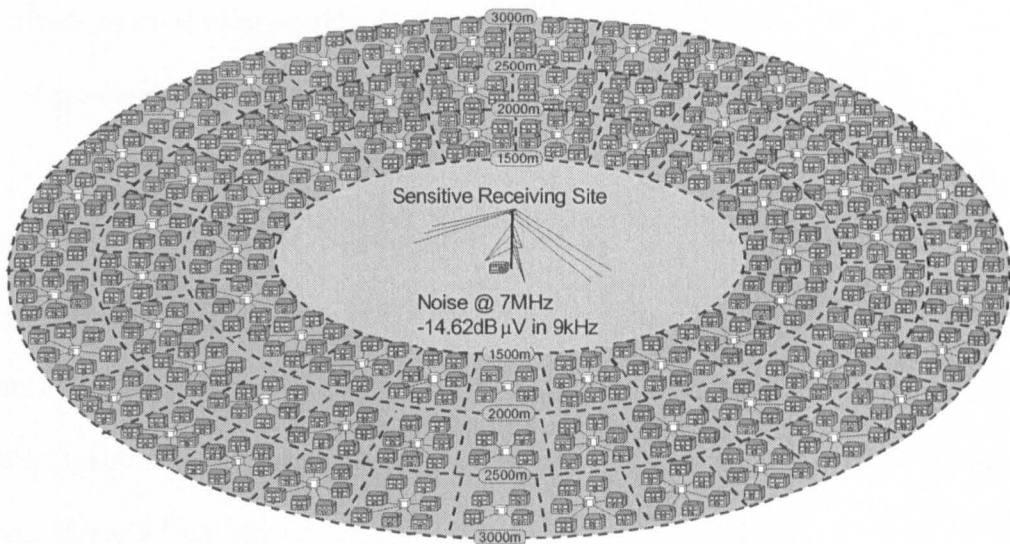


Figure 45 – Noise contribution from multiple PLT cells

In assessing the likely cumulative effect of many PLC ‘cells’ over distances of 100s of meters, it is necessary to establish two key sets of parameters. Firstly - a true indication of

‘launched’ electromagnetic energy for a given PSD. Secondly - an accurate model for field strength regression. Both of these elements have been supplied by this experimental program. [43]

6.2 Mitigation of localised interference by application of EMC Standards

The conflicting interests of PLC manufacturers / network operators and radio users are manifested in the different type of PLC emission standard desired by each party.

Radio operators, seeking to minimise interference to existing radio users, have sought a ‘network compliance’ standard. By specifying a maximum field strength at a given frequency and distance from the power network, such a standard could be used to resolve individual instances of interference. Although PLC standards of this type have typically been derived from similar standards covering (co-axial) cable TV networks, there are a number of practical difficulties when translated into the PLC environment.

The intention with PLC deployment is to make use of the existing power network in its existing state. Metallic shielding, routinely used on cable TV networks cannot be retro-fitted so the only practical means of reducing radiated interference from PLC networks is to reduce the power spectral density of the injected signals. In order to accurately measure such emissions, above a ‘real world’ noise floor, it is necessary to measure over only a very short distance (a few meters). This distance is within the ‘near field’, typically quoted as $\lambda / 2\pi$. Within this region the observed electrical and magnetic fields do not represent true launched electromagnetic energy, but are dominated by magnetic and electrostatic coupling. It could be argued that to a radio receiver or receiving antenna located within the near field of a PLC

network, the distinction is irrelevant. As soon as field strength extrapolations are made to greater distances, this distinction becomes critical.

The pervasiveness of a PLC enabled power network would also necessitate a short measurement distance, since moving the measurement point away from one well defined part of the network usually moves it nearer to other parts of the same network.

PLC manufacturers and potential network operators desire to have an 'equipment compliance' standard. This type of standard would specify maximum injected power spectral density across a given frequency range into a reference mains network similar to that used in CISPR 22. This type of standard would allow PLC modem designs to be tested in an EMC lab and to be CE marked or similar. If a suitably low PSD can be specified, and sensitive frequencies avoided, interference problems could be completely avoided. There is an obvious trade off between maximising data bandwidth and/or signal reach and minimising injected PSD. There is therefore a desire among PLC manufacturers that this limit be as high as possible. The varied design of low voltage power networks throughout the world makes it very difficult to specify a limit that guarantees no interference in all situations.

6.3 Mitigation of cumulative interference by the application of EMC standards.

Cumulative interference from distant PLC systems manifests itself as a potential increase in the existing noise floor. Since almost every part of the radio spectrum has licensed users, it is not possible to eliminate cumulative interference by frequency selection.

It is therefore necessary to estimate the field strength increase caused by a number of PLC installations. The non coherence of these signal sources means that an addition of their received signal power levels rather than detectable field strengths yields more valid results.

By considering likely propagation modes (line of sight / ground wave / sky wave) for the frequency bands used can a suitable regression factor be estimated. For line of sight and sky wave, a 'worst case' regression factor of -20 dB per decade is appropriate, although in the case of sky wave this is over a very large distance. [15]

From the regression characteristics observed in this experimental program, it is clear that the -20 dB per decade 'free space' regression is not valid for ground wave and similar propagation. From the limited number of frequencies examined, it is suggested that a figure of at least -30 dB per decade is valid in the lower part of the HF spectrum. Beyond 10 MHz, where admittedly other propagation modes pre-dominate, even higher figures are appropriate.

Any cumulative mitigation interference standard would have to be based on an 'acceptable' increase in the radio noise floor for some radio users. Clearly, there will have to be a great deal of debate as to what noise increase is acceptable within each part of the HF spectrum - such a debate is beyond the scope of this thesis. It is, however, only after this debate has taken place that the regression figure established during this experimental program, along

with perhaps other similar surveys covering many more locations internationally and across a wider range of frequencies, becomes useful.

It should then become possible, using a theoretical model relating cumulative noise floor increase at a distant point, to measured field strength from individual cells at 3 m or similar, to derive an acceptable field strength figure at 3 m that mitigates cumulative interference to an acceptable level. A similar model for quantifying an acceptable 'exclusion zone' for PLC deployment around a sensitive HF receiving site has already been co-developed by the author and is detailed in Appendix 3 [43].

6.4 Orthogonal Frequency Division Multiplexing (OFDM)

When considering the deployment of a PLC network, it has often been necessary to undertake a frequency response survey. Based on a number of points or nodes on a representative power network, this survey attempts to identify one or more frequency bands where attenuation and noise are minimised. i.e where two way communication with sufficient signal to noise ratio may be possible. Combining this information with prior knowledge for radio frequency usage in the region it possible to identify one or more frequency bands as being suitable for PLC modem operation.

From the above known parameters, Shannon's equation provides a figure for the theoretical maximum data bandwidth. Where only traditional single (modulated) carrier systems are considered the data bandwidth is limited a small fraction of the theoretical maximum by two factors almost unique to PLC.

Firstly, due to multiple signal reflections within the power network, the phase response of the network is highly non-linear across frequency ranges in the order of several hundred kHz and above. Since the exact configuration of each network varies, so this profile of each network is subtly different in each case. Secondly, the noise measured on power networks is location dependant and is often time variant. Orthogonal Frequency Division Multiplexing (OFDM), as discussed in chapter 1 provides an ideal solution that addresses these problems.

An OFDM system uses multiple orthogonal carriers each occupying only a few kHz. Frequency and phase response across such a small frequency band is relatively flat. During the initial setup phase between a pair of OFDM modems, a substantial number of carriers will be affected by attenuation and/or noise and their SNRs will suffer accordingly. With the remaining carriers able to support a robust error detection / correction / re-transmission scheme, an OFDM system can identify those carriers with low SNRs and either raise the transmitted power or abandon carrier transmission as appropriate. The orthogonal nature of the transmitted carriers allows such manipulation on a carrier by carrier basis, such that maximum use can be made of available frequency bandwidth.

Each modem can store the optimised carrier selection and transmitted power levels for every other modem on the network. This optimisation can be dynamically adapted as attenuation and noise on the power network change.

6.5 Interference mitigation by application of OFDM

The same characteristics that make OFDM such a useful technique for adapting to dynamic non-ideal networks, can also be used to implement some of the interference mitigation measures discussed earlier in the chapter.

Typical OFDM implementations are now based on high speed digital signal processing (DSP) systems, where the modulation / demodulation and carrier arrangement is undertaken in software. With such systems it is easily possible to apply a frequency 'mask' to each modem after the time of manufacture. This would allow the same hardware design to be quickly and cheaply modified to operate in a range of different countries / regions with different power network configurations and frequency band plans.

Having masked out sensitive frequency bands, including emergency frequencies, amateur bands and HF broadcast bands, in order to mitigate the potential for localised interference, OFDM allows further steps to be taken to mitigate cumulative interference over larger distances. Instead of optimising the remaining carriers for maximum SNR, and therefore maximum data bandwidth, it is possible to optimise for minimum transmission power spectral density (PSD). The limiting factor for this optimisation could now be a minimum acceptable data bandwidth, either specified at manufacture or calculated dynamically according to the number of concurrent users and/or current data traffic.

6.6 Further work in the field of PLT interference mitigation.

It is the author's opinion that the interference mitigation measures detailed so far in this chapter would provide a sufficiently robust basis to permit wide scale PLC deployment. For the most part, these measures are based on 'worst case' predictions of interference and as such may be overly restrictive. This would translate directly into a reduced data bandwidth available to PLC network operators and thus a less attractive commercial proposition. In fact a number of international standards bodies are continuing to gather data on field strength and

radiated signal power from commercial PLT systems, and this information will prove invaluable in finalising the limits specified in their standards.

It has been suggested that, in addition to the fixed frequency mask, it might be possible to allow transmission on certain frequencies on a dynamic basis. For HF frequencies licensed for two way communication, the local transmissions are likely to be detected as noise on one or more PLC channels. The non ideal nature of the power network as a HF antenna means that the distant transmission is very unlikely to be detectable.

By detecting the local transmission it would be possible for the PLC modem to identify which frequency bands were required for use and immediately cease transmission on those frequencies until a specified period of time after the last detectable transmission. In a further variation on this system, it could be arranged for the licensed radio operator to 'control' PLC transmissions on a radio band by transmitting simple CW/Morse sequences to the PLC modem.

7 Chapter 7 – Summary of the Main Points of Research Program

From this research program it has been possible to demonstrate a number of key points, with respect to the potential deployment of power line communications on LV distribution networks.

The majority of UK and European LV networks currently in use are based on what might be called modern specifications and installation practices - those employed throughout the past 50 years. Whilst non ideal from a radio frequency perspective, a typical LV network does provide access to upwards of 100 potential communications customers. In fact between 200 and 300 customers are typical. Although not demonstrated in this thesis, others have demonstrated [44] that it would be possible to build a viable business model around wide scale deployment of such systems.

Since LV networks were designed exclusively for the delivery of electrical power at 50 / 60 Hz, their RF performance was never previously considered. For any given LV network and chosen communications frequency, the attenuation between two communication nodes, along with the radiated emissions caused by transmission from any single node are fixed. It is possible to improve the RF performance of a network by the application at strategic points of RF filters [45] but this cost must be supported by the business model. Given these constraints, there are two criteria that a power line communication system must meet.

Firstly, the communications path between two nodes must offer a sufficiently high signal to noise ratio and therefore data bandwidth to enable a competitive service to be offered. With the type of modulation schemes likely to be employed, this requires the power spectral

density to be high as possible. It is also necessary to remember that whatever bandwidth is made available, it is contended between all telecoms customers on that substation.

Secondly, the radiated emissions from each transmitting node will depend entirely on the structure of the LV network, in particular the exact design of the LV network in the area local to the transmitting node. Factors such as overhead conductors, street furniture (street lights etc.) and cable types will all influence the degree of radiated emissions, and largely cannot be changed. Therefore the only means by which the radiated emissions can be kept within an 'acceptable' level, the definition of acceptable will be discussed below, is to minimise the transmitted power spectral density. It is this requirement which must be prioritised. Analysis of these results, by the author of this thesis has demonstrated that a launch PLT power spectral density of -60 dBm/Hz, measured in 9 kHz bandwidth, should be compatible with the requirement for non interference with existing radio users (with the exception of emergency frequencies). For emergency frequencies it would be necessary to employ frequency masking.

Turning to the topic of radiated field strength regression, it has been shown that the observed regression curves, studied at multiple frequencies, exhibit a fairly constant -35 dB/decade regression across both rural and semi-urban environments. The figures also show a good correlation with the rough ground theoretical model with the land usage factor set to 10%, that of a rural environment. Other models, including the flat earth model, EGLI and CCIR models were also examined in chapter 5 but proved the correlation was less clear.

International EMC standards, either PLT specific, or extensions of existing standards, are critical to the commercial manufacture and deployment of PLT. There must be two distinct

standards groups. Firstly, an 'equipment compliance' standard will specify permitted differential and common mode conducted emissions into a test network (as specified in CISPR22). This standard would specify a power spectral density in dBm/Hz in a 9 kHz measurement bandwidth. Since PLT systems are emerging that use frequencies up to 80 MHz, it would be wise for the standard to cover frequencies up to 100 MHz. Based on the experimental work and results in this thesis, a figure of -60 dBm/Hz is proposed, excepting parts of the spectrum used for emergency / safety of life radio services

Given the variation in components and layout of each power network, a different RF performance, and hence antenna factor, can be expected from every transmitting point on each network. Inevitably there will be some examples which exhibit a higher radiated emission than would conventionally be expected. In order to meet the non interference requirement in these situations, a second interference mitigation standard is required, along the same lines as the radiated emission part of CISPR 22. This would allow interference problems on specific networks to be resolved. Such a standard is currently under development in a collaborative process involving CENELEC, specifically the European EMC committee for radiated emissions from wireline networks, and the world committee of CISPR 22.

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